

A LOW-NOISE 100-MC BANDWIDTH TRANSISTORIZED
I-F AMPLIFIER FOR RADIO ASTRONOMY

APPROVED:

A LOW-NOISE 100-MC BANDWIDTH TRANSISTORIZED
I-F AMPLIFIER FOR RADIO ASTRONOMY

by

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THESIS

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PREFACE

This thesis describes the development of a low-noise wideband I-F amplifier suitable for research in radio astronomy. The following work was supported in part by the National Aeronautics and Space Administration Grant NSG 432 and was performed at the Electrical Engineering Research Laboratory of The University of Texas.

To Mr. C. W. Tolbert and Dr. A. W. Straiton of EERL goes my sincere appreciation for their invaluable guidance and encouragement. Also I would like to thank all of the personnel of EERL who contributed to the preparation of this thesis.

Ronald Alfred Vivian

Austin, Texas

July, 1964

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I. INTRODUCTION

Radio astronomy is a recent branch of astronomy concerned with the study of the universe by means of radio techniques. The necessity for specialized radio astronomy instrumentation, capable of detecting very low-power signals from radio sources, has led to the development of high-gain directive antennas and low-noise receivers. This report concerns the design and construction of a transistorized amplifier suitable for use in radio astronomy or other applications requiring a low-noise wideband amplifier.

A. Radiometers Used in Radio Astronomy

The electromagnetic radiation of hot bodies is a fundamental property of matter. Since the power density of the signal received from the thermal radiation of a body is proportional to its temperature, radio astronomers commonly express the sensitivity of the receiving system in terms of the minimum temperature change, in degrees Kelvin, which can be detected.¹ Thermal radiation is an essentially incoherent radiation having the statistical characteristics of noise, similar to that generated in receivers.^{2,3} Consequently, any noise generated in the receiver or random changes in the power gain of the system will lead to error when very low-level noise signals are received.

The sensitivity of a simple radio telescope, which records the total output power of the receiver, is limited because any gain change in the receiver shows up directly in the recorded output. The development

of radiometers using special comparison switching techniques has circumvented this strong dependence on receiver gain stability. In the switched input radiometer, first described by Dicke⁴, the input signal to the receiver is modulated by a switch driven at a low frequency. The receiver input is switched between a reference temperature source and the received noise temperature at the antenna. The difference in the two temperatures appears as modulation of the receiver power level and is converted into a dc signal by synchronously detecting the video output of the receiver.

The major advantage of the switched input radiometer lies in the fact that it is not subject to large changes in the output level as a result of gain variations of the receiver. In the unswitched type of radiometer, a change in output level equal to the percentage gain change times the input noise temperature is observed. In contrast, the switched input radiometer exhibits a change in output level equal to the percentage gain change times the difference in input noise temperature and reference temperature.

B. Radiometer Sensitivity

In addition to gain changes, the radiometer output level is also a function of receiver bandwidth, noise figure, and output integration time. A generalized expression for the minimum detectable noise signal temperature for the switched input radiometer can be given by³:

$$\Delta T = K \left[\frac{(F - 1) T_o}{\sqrt{B \tau}} + \frac{(G_t - G_o)}{G_o} (T_r - T_d + T_a) \right] \quad (1)$$

where

- ΔT = the minimum detectable temperature change
- K = a system constant
- F = the system noise figure
- T_o = reference ambient temperature, 290°K
- B = the predetection bandwidth
- τ = output integration time
- G_t = the gain of the receiving system at time t
- G_o = the average value of G_t over τ
- T_r = the reference temperature observed by the receiver
- T_d = antenna terminal temperature observed by the receiver, and
- T_a = the effective antenna temperature produced by the source.

C. Amplifier Requirements

The radiometer in present use at the Electrical Engineering Research Laboratory of The University of Texas has a switched input millimeter wavelength crystal mixer superheterodyne receiver. An I-F amplifier having a center frequency of 30 mc or 60 mc and bandwidth of about 10 mc is used, along with a synchronous detector. From Equation (1), it can be seen that the system can be considerably improved by increasing the I-F amplifier bandwidth and output integration time. As a result, a transistorized amplifier with a bandwidth of 100 mc, a low noise figure, and a power gain of 70 to 80 db was proposed. Gain stability with respect to time, temperature, and supply voltage variations were also prime considerations in the proposed amplifier. The present radiometer uses a balanced crystal mixer, in the form of a hybrid waveguide junction, known

as a "Magic T," which produces an improvement in noise figure over the conventional single-ended crystal mixer.⁵ Therefore, the amplifier should have a balanced input impedance of 300 to 600 ohms, depending on the crystal mixer biasing. The output impedance of the amplifier should be suitable for a high impedance video detector.

II. WIDEBAND AMPLIFIERS

Several methods can be used to obtain large bandwidths in transistorized amplifiers. The most common methods used are stagger-tuning of stages, frequency response shaping by compensation networks, broadband interstage transformers, and frequency-dependent feedback.

A. Transistor Characteristics

Transistors designed for use in VHF and UHF applications have a reasonably low noise figure which is nearly constant from about 1 mc to above 100 mc, gradually increasing at higher frequencies. This noise figure characteristic indicates that a low center frequency should be used in the proposed transistorized amplifier. However, the low-frequency edge of the passband should be above the highly congested high-frequency bands. In view of these factors, a center frequency of 100 mc was chosen for the amplifier.

The wide passband of 100 mc and the relatively low center frequency of 100 mc seem to rule out the use of the stagger-tuned-stage amplifier configuration, since a very low tuned circuit Q and a large number of stages would be required.

The common-emitter and the common-base amplifier configurations offer the best characteristics for high power gain. However, both present difficulties in obtaining the required constant power gain over the 100-mc bandwidth. Above f_{hfe} , the short-circuit emitter cutoff frequency (about 30 mc in most UHF transistors), the common-emitter amplifier current

gain decreases at a rate of about 6 db per octave up to f_T , the frequency at which the short-circuit current gain is 0 db. This latter frequency is commonly called the gain-bandwidth product.

The common-base transistor amplifier configuration exhibits a nearly constant forward current gain, slightly less than unity, up to f_{hfb} , the short-circuit common-base current gain cutoff frequency. Most UHF transistors have an f_{hfb} of several hundred megacycles. However, the common-base amplifier output impedance must be greater than the input impedance for any power gain realization.

B. Amplifiers Using Wideband Transformers

Bell Telephone Laboratories have succeeded in designing some miniature broadband toroid transformers for use in common-base transistorized amplifiers having 20-to 50-mc bandwidths centered at 75 mc.⁶⁻⁸ These transformers are of the transmission line type and have a turns ratio of 2:1, which gives a 4:1 impedance transformation and, consequently, a power gain of about 6 db per stage. Sufficient design data was not available for duplication of the wideband transformers.

C. High-Frequency Compensation Networks

Some shaping of gain-frequency characteristics of the common-emitter transistor amplifier is possible by using high-frequency compensation or equalization networks. A network with a decreasing attenuation of 6 db per octave over the desired bandwidth can be used to offset the current gain falloff of the transistor. However, this method of gain-frequency shaping

becomes increasingly difficult when used for large bandwidths at low frequencies. The Bell Telephone Laboratories have used this technique for low-noise input stages in I-F amplifiers with bandwidths of up to 50 mc.^{8,9}

D. Amplifiers with Frequency-Dependent Feedback

Perhaps the best way to obtain large bandwidths is by using negative feedback to reduce the low-frequency gain of the common-emitter transistor amplifier and removing the feedback at higher frequencies. This can be accomplished by using a frequency-dependent shunt feedback network in the form of a series R-L network between the collector and base of the transistor amplifier.^{9,10} The low-frequency power gain is determined by the series resistor, and the inductive reactance effectively removes the feedback at high frequencies. This low-pass amplifier configuration is quite flexible in that gain and bandwidth can easily be exchanged, with the resultant gain-bandwidth product approximately given by f_T . In addition, input impedance is decreased and can be made nearly constant by proper choice of the feedback network. A high-pass filter can be incorporated in the amplifier to attenuate frequencies below 50 mc.

III. SHUNT FEEDBACK AMPLIFIER DESIGN

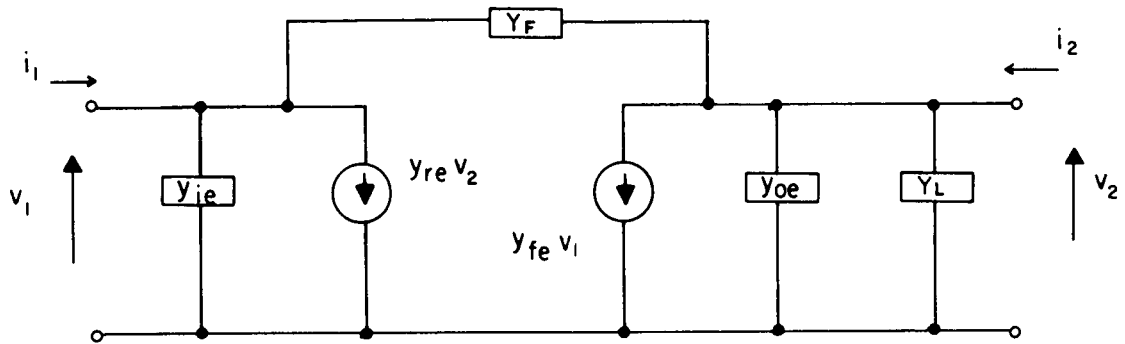
Since the overall noise figure of an amplifier is essentially determined by the noise figure and gain of the input stages, more expensive low-noise transistors with a large gain-bandwidth product, f_T , are desirable in the front end of the amplifier. Silicon transistors were preferred for the amplifier because of their higher temperature operating limits. The 2N2857 silicon NPN transistor, with a 1000-mc f_T and a low noise figure, was chosen for the first two stages of the amplifier. A less expensive transistor, the 2N2708, also a silicon NPN, was chosen for the remaining stages. This transistor has a minimum f_T of 700 mc. Manufacturer's specification sheets for the two transistors are shown in Appendices 1 and 2.

A. Small-Signal Model

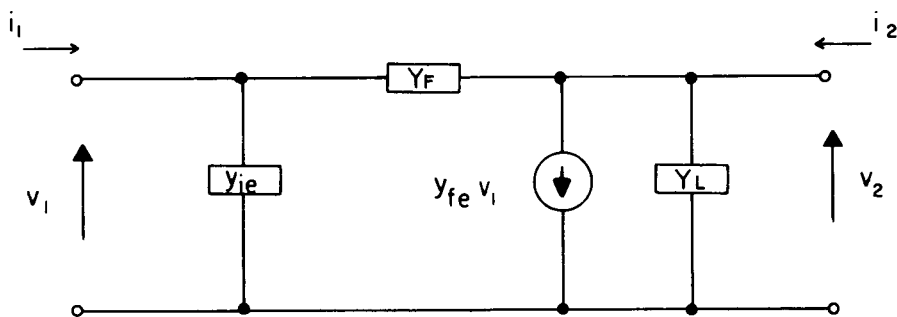
By using negative shunt feedback, the operating characteristics of the common-emitter transistor amplifier can be considerably modified. The small-signal admittance parameter transistor model, shown in Fig. 1-a, will be used to show the effect of negative shunt feedback. The current generator in the input, $y_{re} v_2$, is the result of internal feedback in the transistor and may be neglected, since the external feedback will be much greater. In addition, the output admittance, y_{oe} , is much smaller than the load admittance, Y_L , and also can be neglected. The small-signal admittance parameter model can be considerably simplified, as shown in Fig. 1-b.

B. Voltage and Current Gain Characteristics

The following equations may be written for the circuit in Fig. 1-b:



ADMITTANCE PARAMETER EQUIVALENT CIRCUIT
(a)



SIMPLIFIED EQUIVALENT CIRCUIT
(b)

SMALL-SIGNAL MODELS

FIG. 1.

$$(y_{ie} + Y_F)v_1 - Y_F v_2 = i_1 \quad (2)$$

$$(y_{fe} - Y_F)v_1 + (Y_L + Y_F)v_2 = 0 \quad (3)$$

The voltage gain, A_v , for the circuit can be obtained from Eq. (3).

$$A_v = \frac{v_2}{v_1} = - \frac{(y_{fe} - Y_F)}{(Y_L + Y_F)} \quad (4)$$

Expressions for the current gain, A_i , in terms of the admittance parameters, input and output admittances, and h_{fe} , the short-circuit common-emitter current gain, can be given by:

$$A_i = \frac{i_2}{i_1} = - A_v \frac{Y_L}{Y_i} = h_{fe} \quad (5)$$

where

$$y_{fe} = y_{ie} h_{fe} \quad (6)$$

Since the input admittance, Y_i , of one stage is the output load admittance, Y_L , of the preceeding stage, the interstage current gain has the same magnitude as the voltage gain.

The effect of the shunt feedback network on the current gain can be found from Equations (2), (5), and (6). The current gain for the cases of no feedback, resistive feedback, and series R-L feedback is shown

qualitatively in Fig. 2-a.^{10, 11} It can readily be seen that the low-frequency current gain is decreased and the high-frequency current gain increased by the series R-L feedback network,

C. Input Impedance Characteristics

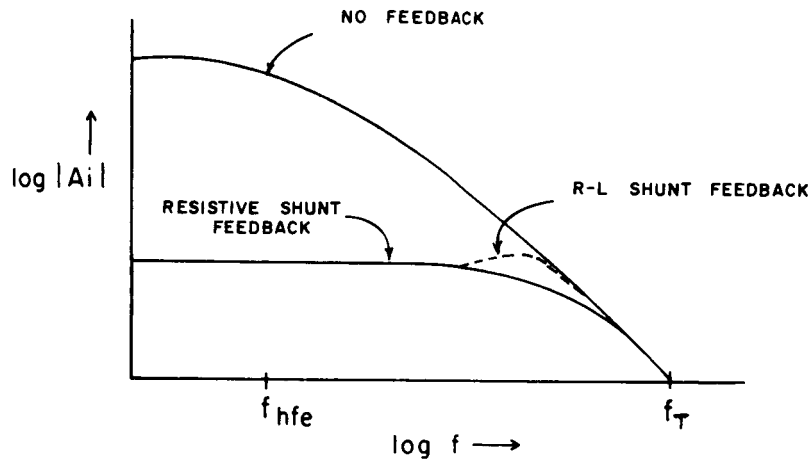
The input admittance of the equivalent circuit shown in Fig. 1-b may be found by combining Equations (2), (3), and (4).

$$Y_i = \frac{1}{Z_i} = \frac{i_1}{v_1} = y_{ie} + Y_F(1 - A_v) . \quad (7)$$

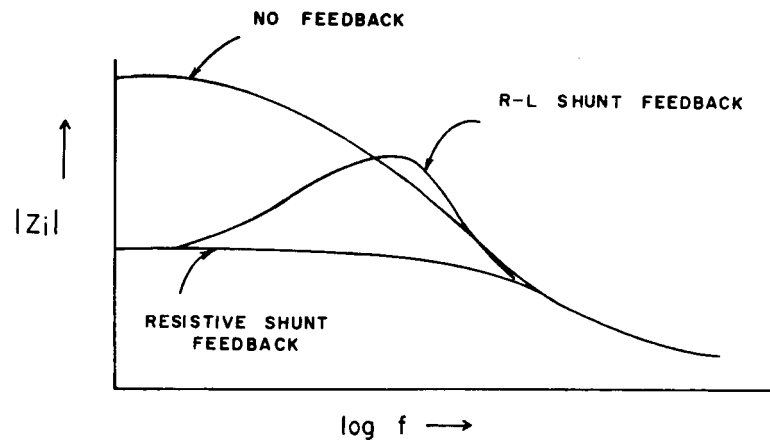
Remembering that A_v is a negative quantity, the input admittance is essentially determined by the voltage gain and feedback admittance, since y_{ie} is a relatively small quantity. The effect of negative shunt feedback on the input impedance is shown qualitatively in Fig. 2-b.¹¹ The input impedance peaking with the series R-L feedback network is caused by the resonant effect of the transistor input capacitance and the feedback inductance. The input stage of the amplifier should have resistive shunt feedback if a nearly constant input impedance is desired.

D. Feedback Network Component Values

An estimate of the component values of the R-L shunt feedback network may be made by first computing the approximate gain per stage of the amplifier. The 2N2708 transistor has an f_T of about 700 mc. For an upper cutoff-frequency of 150 mc, the current gain of a single stage will be 4.7, or 13.4 db. However, the gain loss due to input



CURRENT GAIN VS. FREQUENCY
(a)



INPUT IMPEDANCE VS. FREQUENCY
(b)

CURRENT GAIN AND INPUT IMPEDANCE CHARACTERISTICS

FIG. 2.

and output impedance mismatching is high, so that in practice a gain of about 3, or 10 db, per stage can be expected. To simplify calculations, the magnitude of h_{fe} , which is decreasing at about 6 db per octave over the frequency range of interest, will be used.

By combining Equations (4) through (6), an expression relating the feedback admittance, desired low-frequency current gain, and the transistor parameters may be obtained.

$$Y_F = \frac{y_{ie} (h_{fe} - A_i)}{1 + A_i \left[1 + \frac{y_{ie}}{Y_L} (1 + h_{fe}) \right]} \quad (8)$$

The magnitude of h_{fe} at the low-frequency edge of the passband can be found by dividing f_T by 50 mc. Using the resulting value of 14 for h_{fe} , a load impedance of 50 ohms, a current gain of 3, and the magnitude of y_{ie} from the specification sheet of Appendix 1, the feedback admittance, Y_F , is found to be about 3.7 millimhos, or 270 ohms.

The high-frequency response of the amplifier may be increased by peaking the input impedance slightly below 150 mc. This can be accomplished by setting the inductive reactance of the feedback network equal to the capacitive reactance of y_{ie} . From the specification sheet of the 2N2708, the reactive part of y_{ie} is about +3 millimhos at 150 mc. From Equation (7), it is found that a feedback inductor value of about 1 microhenry is necessary to cancel the capacitive reactance of y_{ie} . However, circuit capacitance was

not included in the calculation, so the actual value of feedback inductance will be considerably smaller than the calculated value.

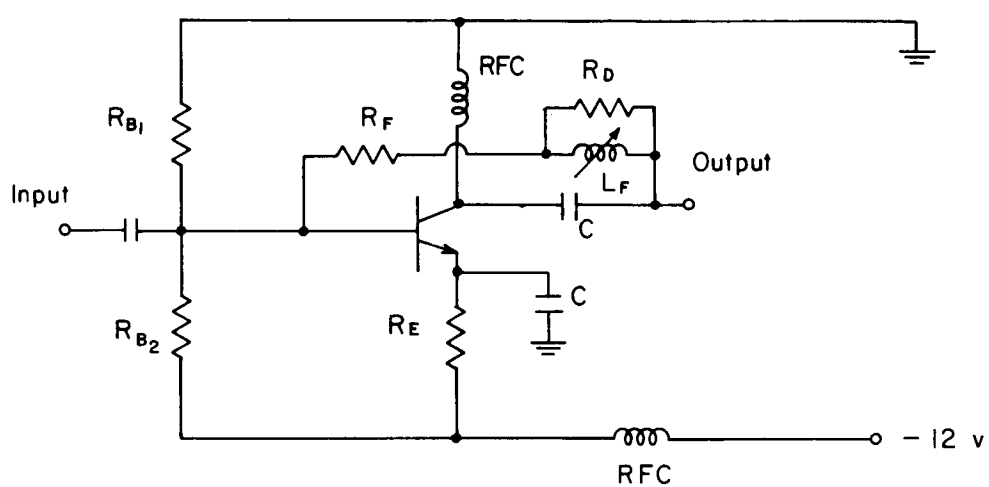
A resistor shunted across the feedback inductor, as shown in Fig. 3, is used to damp the resonant effect of the inductor and input capacitance. Thus, a wide range of gain control over the amplifier passband is available by proper choice of feedback network component values.

E. Transistor Biasing

The next problem is to calculate the values of the biasing resistors for the transistors. The two input stages should have a dc quiescent point which optimizes the noise figure of the transistor. From the specification sheet of the 2N2857 in Appendix 2, the dc points consistent with a low noise figure are $V_{CE} = 6$ volts and $I_C = 1.5$ ma. The voltage across the emitter biasing resistor, R_E , should be large enough to swamp out V_{BE} , which is about 600 mv for silicon transistors. Assuming a supply voltage of 12 volts, the dc voltage across R_E can be set at 6 volts by using a radio-frequency choke from collector to ground. For a collector current of 1.5 ma, R_E should then be 3900 ohms.

The base biasing voltage, which should be 6.6 volts, is supplied by a resistive voltage divider, R_{B1} and R_{B2} , as shown in Fig. 3. Values of 4700 ohms and 5600 ohms are chosen for these two resistors.

The dc operating point for the interstage transistors should optimize the power gain. The specification sheet of the 2N2708 in Appendix I



TYPICAL SHUNT FEEDBACK
AMPLIFIER STAGE

FIG. 3.

indicates that h_{fe} is within 5% of its maximum value with $I_C = 5 \text{ ma.}$ V_{CE} is again set at 6 volts and, consequently, R_E should be 1200 ohms. The voltage divider resistors for the base biasing voltage have the same values as before.

IV. EXPERIMENTAL DATA

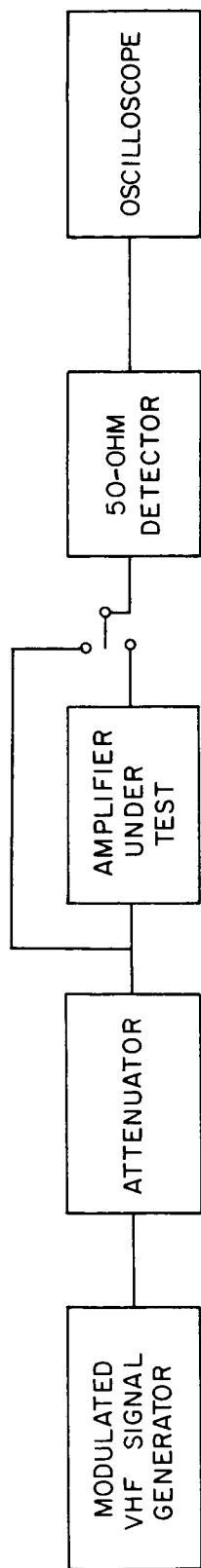
The desired characteristics of the proposed I-F amplifier, previously discussed in the initial pages of this report, are:

- (1) bandwidth of 100 mc centered at 100 mc
- (2) low noise figure
- (3) power gain of 70 to 80 db
- (4) flat frequency response
- (5) balanced input impedance of 300 to 600 ohms
- (6) high gain stability, and
- (7) output suitable for a high impedance video detector.

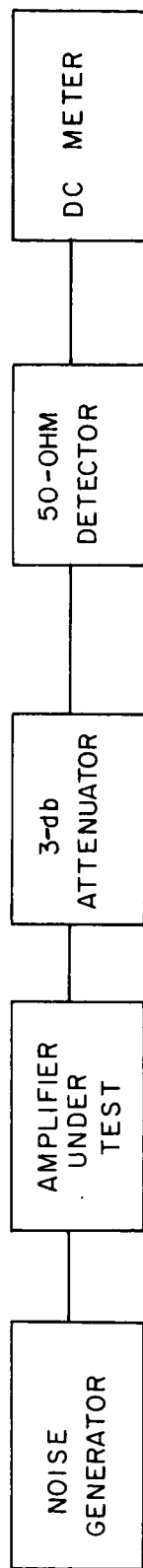
A. Test Equipment Layout

Block diagrams of the test equipment layout used in building the amplifier are shown in Fig. 4. The signal generator shown in Fig. 4-a was modulated with a 1-kc sine wave, permitting the use of low-frequency measuring instruments. The attenuation network consisted of a 100-db range attenuator, calibrated to 1 db, and a 10-db precision attenuator, calibrated to 2/10 db. Initially, the load impedance of the amplifier was set at 50 ohms. The insertion gain versus frequency of the amplifier was measured by using a video crystal detector and a low-frequency oscilloscope or VTVM.

The block diagram in Fig. 4-b shows the method used in measuring the noise figure of the amplifier. The noise generator used provided either balanced or unbalanced inputs for the amplifier. The



TEST SET-UP FOR MEASURING AMPLIFIER INSERTION GAIN VS. FREQUENCY
(a)



TEST SET-UP FOR MEASURING AMPLIFIER NOISE FIGURE
(b)

TEST EQUIPMENT LAYOUT

FIG. 4.

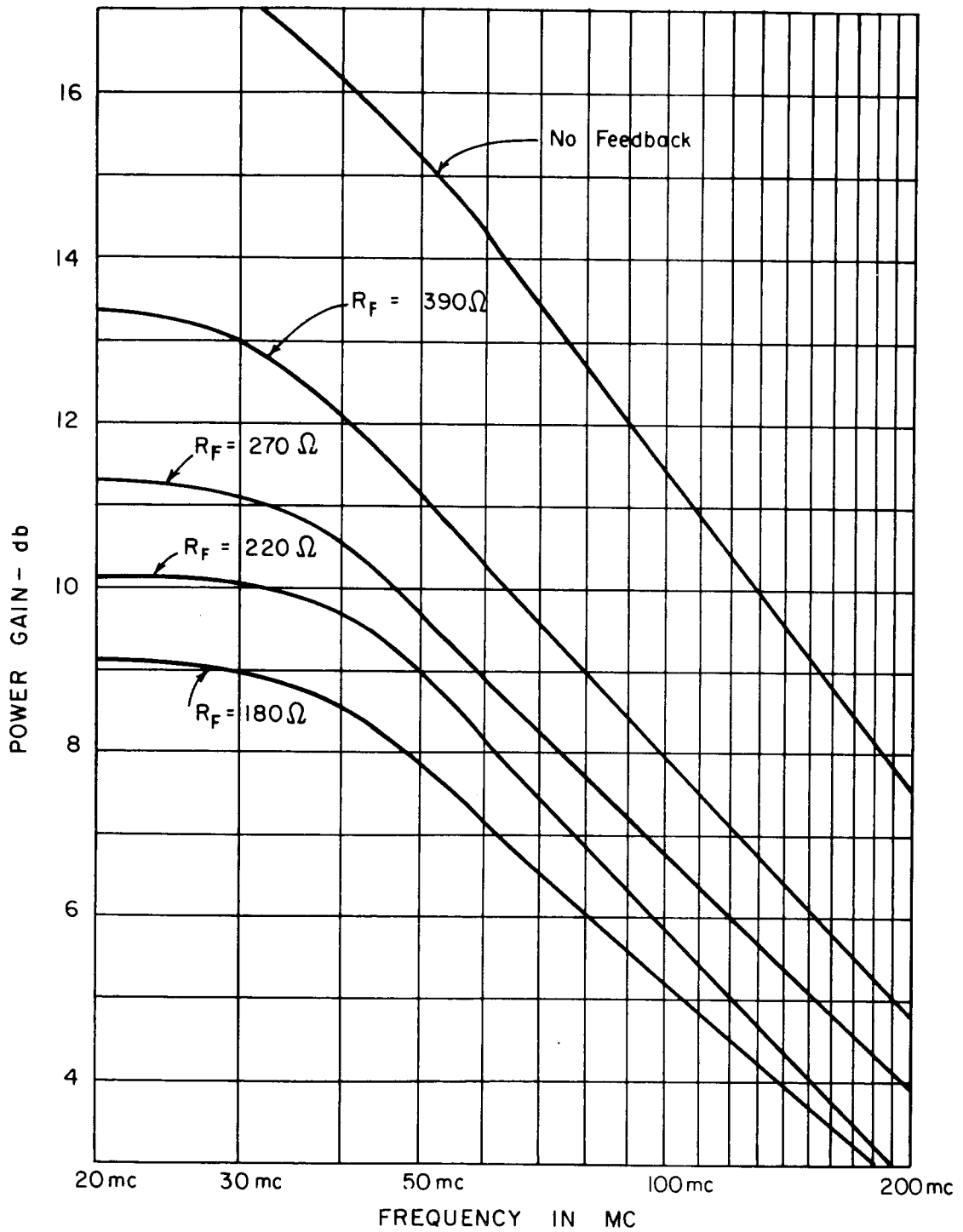
noise figure was found by first noting the dc output voltage of the amplifier, inserting the 3-db pad, and increasing the output level of the noise generator until the dc output voltage returned to its initial value. The noise figure of the amplifier could then be read directly from the noise generator.

B. Single-Stage Amplifier

A single-stage amplifier using the 2N2708 transistor was constructed and insertion gain versus frequency measured for several values of shunt feedback impedance. A circuit diagram of the amplifier is shown in Fig. 3 and the gain-frequency characteristics are shown in Figs. 5 and 6. The peaking effect of the inductors in the feedback network is evident in Fig. 6.

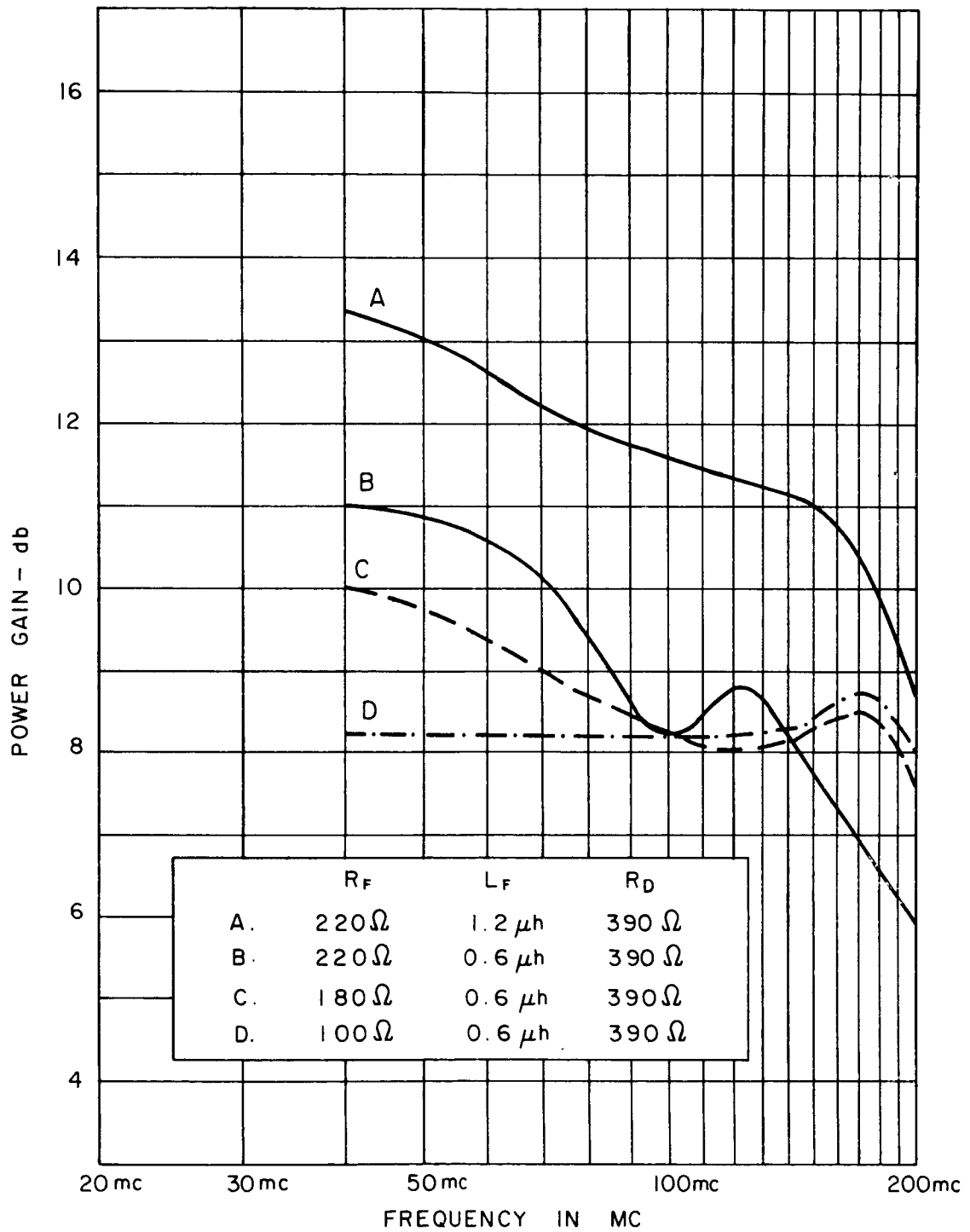
C. Four-Stage Amplifier

Next, a four-stage amplifier using 2N2708 transistors was constructed. A slight gain instability was noted and apparently was due to a low-frequency resonant effect between the collector radio-frequency choke and circuit capacitance. A 1000-ohm resistor placed in parallel with the choke effectively damped the resonance and stabilized the amplifier. In addition, a metal cover over the bottom of the amplifier chassis was necessary to prevent instability from input stage pickup of stray fields. A circuit diagram of the four-stage amplifier is shown in Fig. 7 and the frequency response for various values of feedback impedance is shown in Figs. 8 through 10. The effects of changing one feedback network



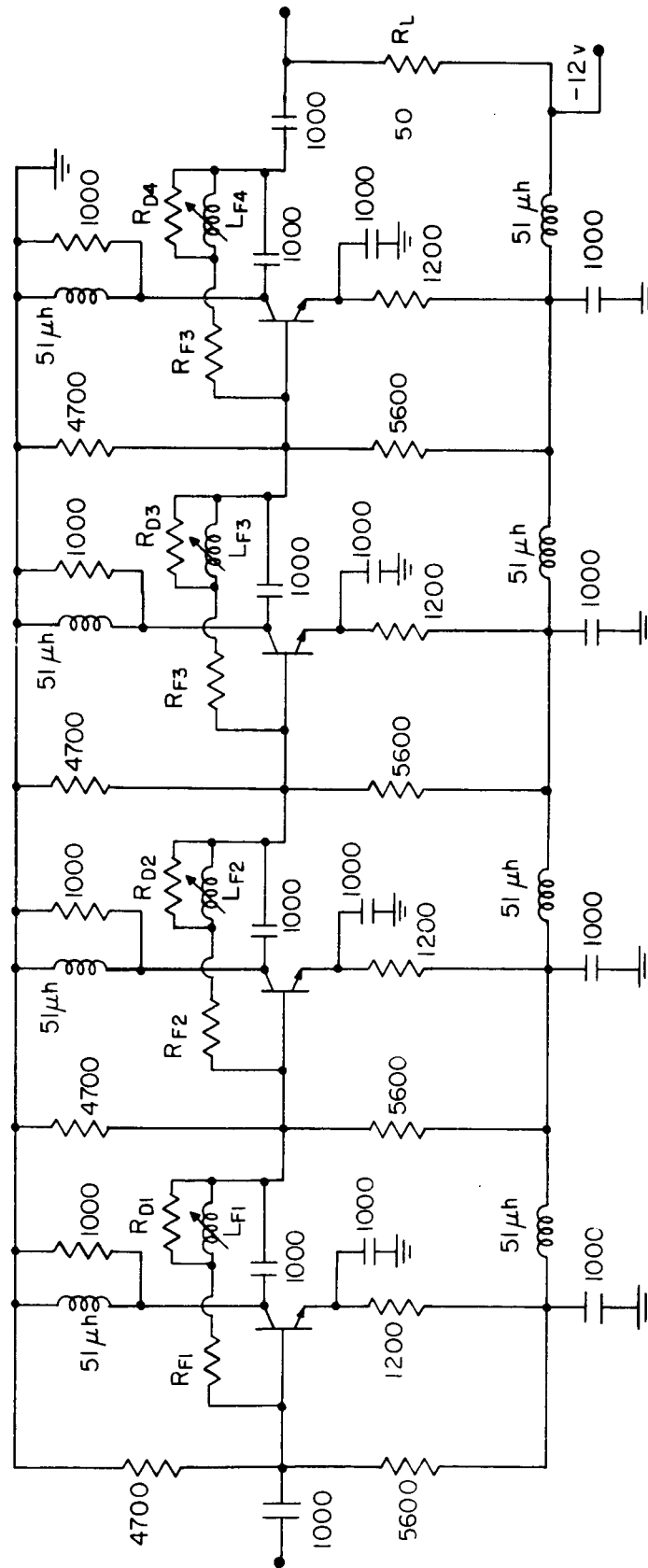
FREQUENCY RESPONSE OF SINGLE-STAGE
AMPLIFIER; RESISTIVE SHUNT FEEDBACK

FIG. 5.



FREQUENCY RESPONSE OF SINGLE-STAGE
AMPLIFIER; R-L SHUNT FEEDBACK

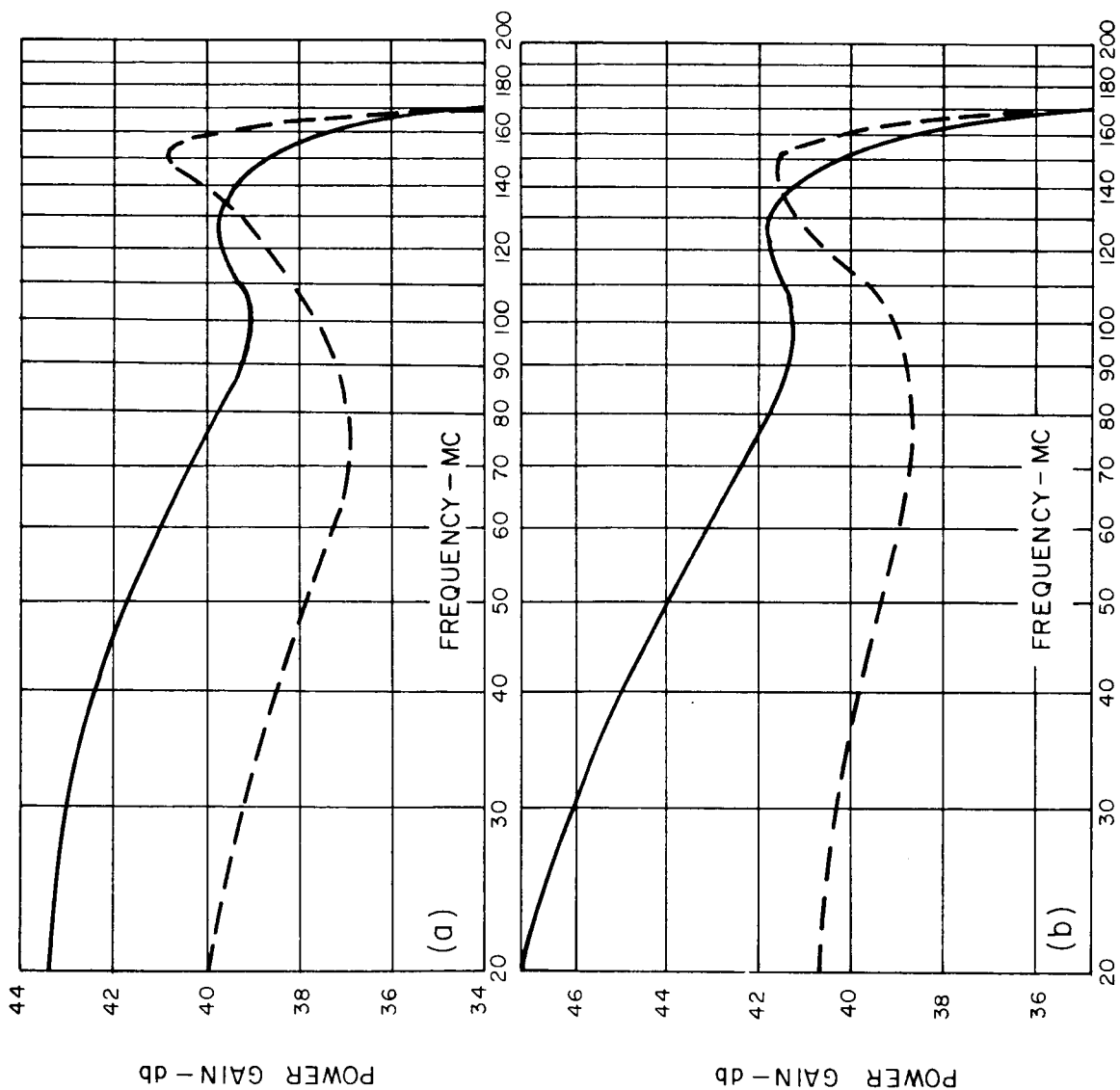
FIG. 6.



All Transistors are 2N2708
All Capacitors in Picofarads

FOUR-STAGE AMPLIFIER CIRCUIT

FIG. 7.



$$\text{---} = R_{F1} = R_{F4} = 220 \Omega$$

$$\text{---} = R_{F2} = R_{F3} = 180 \Omega$$

$$L_{F1}, L_{F3} = 0.6 \mu h$$

$$L_{F2}, L_{F4} = 0.3 \mu h$$

$$R_{D1}, R_{D3} = 390 \Omega$$

$$R_{D2}, R_{D4} = 220 \Omega$$

$$\text{---} = R_{F1} = R_{F4} = 220 \Omega$$

$$\text{---} = \begin{cases} R_{F1}, R_{F3} = 220 \Omega \\ R_{F2}, R_{F4} = 120 \Omega \end{cases}$$

$$L_{F1}, L_{F3} = 0.9 \mu h$$

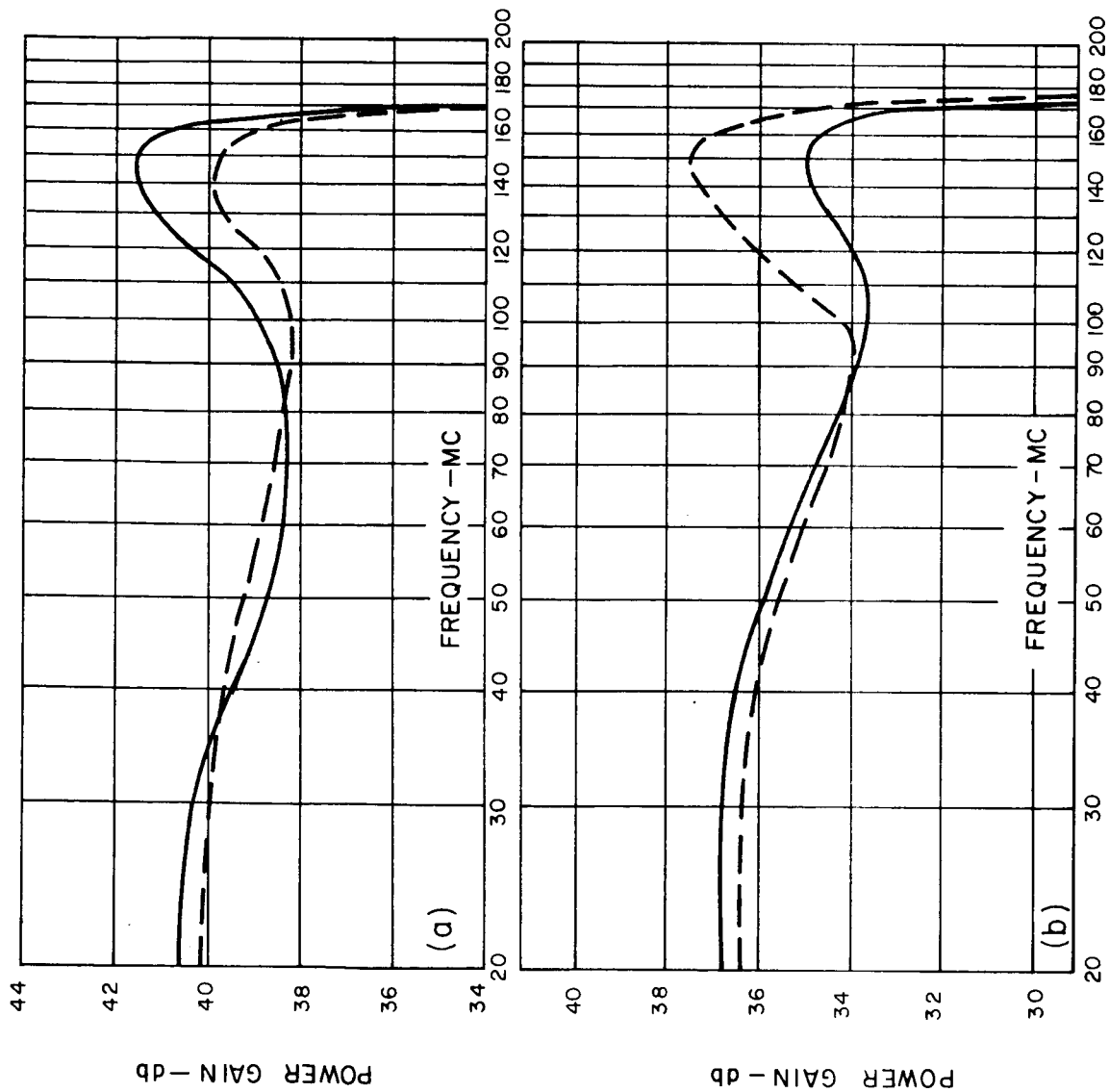
$$L_{F2}, L_{F4} = 0.3 \mu h$$

$$R_{D1}, R_{D3} = 390 \Omega$$

$$R_{D2}, R_{D4} = 220 \Omega$$

FREQUENCY RESPONSE OF FOUR-STAGE AMPLIFIER; EFFECT OF R_F

FIG. 8.



$$\text{---} = \begin{cases} R_{D1}, R_{D3} = 390 \Omega \\ R_{D2}, R_{D4} = 330 \Omega \end{cases}$$

$$\text{---} = \begin{cases} R_{D1}, R_{D3} = 390 \Omega \\ R_{D2}, R_{D4} = 270 \Omega \end{cases}$$

$$L_{F1}, L_{F3} = 0.9 \mu h$$

$$L_{F2}, L_{F4} = 0.3 \mu h$$

$$R_{F1}, R_{F3} = 220 \Omega$$

$$R_{F2}, R_{F4} = 120 \Omega$$

$$\text{---} = \begin{cases} R_{D1}, R_{D3} = 330 \Omega \\ R_{D2}, R_{D4} = 220 \Omega \end{cases}$$

$$\text{---} = R_{D1} - R_{D4} = 220 \Omega$$

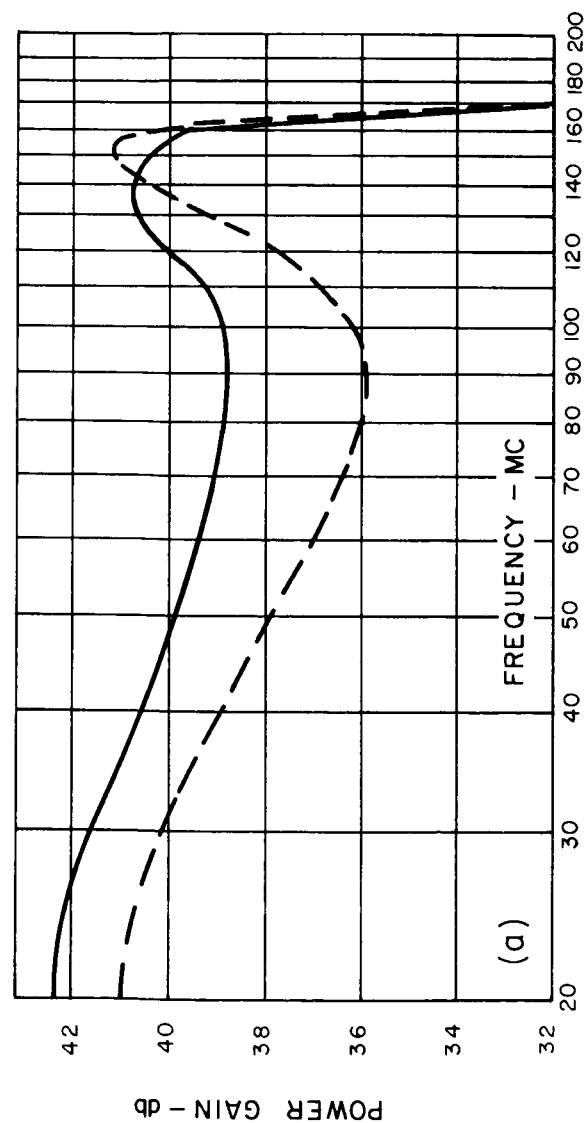
$$L_{F1}, L_{F3} = 0.9 \mu h$$

$$L_{F2}, L_{F4} = 0.3 \mu h$$

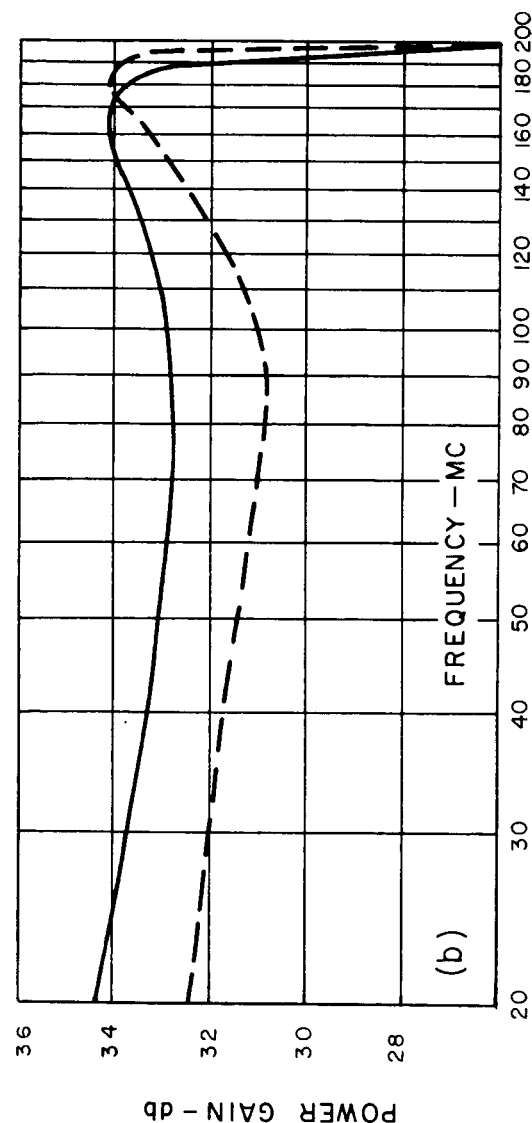
$$R_{F1} - R_{F4} = 220 \Omega$$

FREQUENCY RESPONSE OF FOUR-STAGE AMPLIFIER; EFFECT OF R_D

FIG. 9



$\text{---} = \begin{cases} L_{F1}, L_{F3} = 0.9 \mu h \\ L_{F2}, L_{F4} = 0.3 \mu h \end{cases}$
 $\text{---} = \begin{cases} L_{F1}, L_{F3} = 0.5 \mu h \\ L_{F2}, L_{F4} = 0.3 \mu h \end{cases}$
 $R_{F1}, R_{F3} = 220 \Omega$
 $R_{F2}, R_{F4} = 150 \Omega$
 $R_{D1} - R_{D4} = 330 \Omega$



$\text{---} = \begin{cases} L_{F1}, L_{F3} = 0.5 \mu h \\ L_{F2}, L_{F4} = 0.6 \mu h \end{cases}$
 $\text{---} = \begin{cases} L_{F1}, L_{F3} = 0.5 \mu h \\ L_{F2}, L_{F4} = 0.3 \mu h \end{cases}$
 $R_{F1}, R_{F3} = 70 \Omega$
 $R_{F2}, R_{F4} = 120 \Omega$
 $R_{D1}, R_{D3} = 330 \Omega$
 $R_{D2}, R_{D4} = 150 \Omega$

FREQUENCY RESPONSE OF FOUR-STAGE AMPLIFIER; EFFECT OF L_F
FIG. 10.

component value while holding the other values constant clearly indicate the wide range of gain control available over the passband.

D. Output Saturation by Noise Voltage

The maximum output voltage of the 2N2708 transistor, with shunt feedback from collector to base and a load impedance of 50 ohms, was found to be about 100 millivolts peak-to-peak, due to a combination of low voltage gain, input saturation, and feedback limiting of the collector current swing. The noise voltage generated across the source resistance, R_s , is given by:¹²

$$E_s^2 = 4kTBR_s \quad (9)$$

where

- E_s^2 = the mean square noise voltage
- k = Boltzmann's constant, 1.372×10^{-23} joules per degree Kelvin
- T = temperature in degrees Kelvin
- B = bandwidth in cycles per second, and
- R_s = the source resistance.

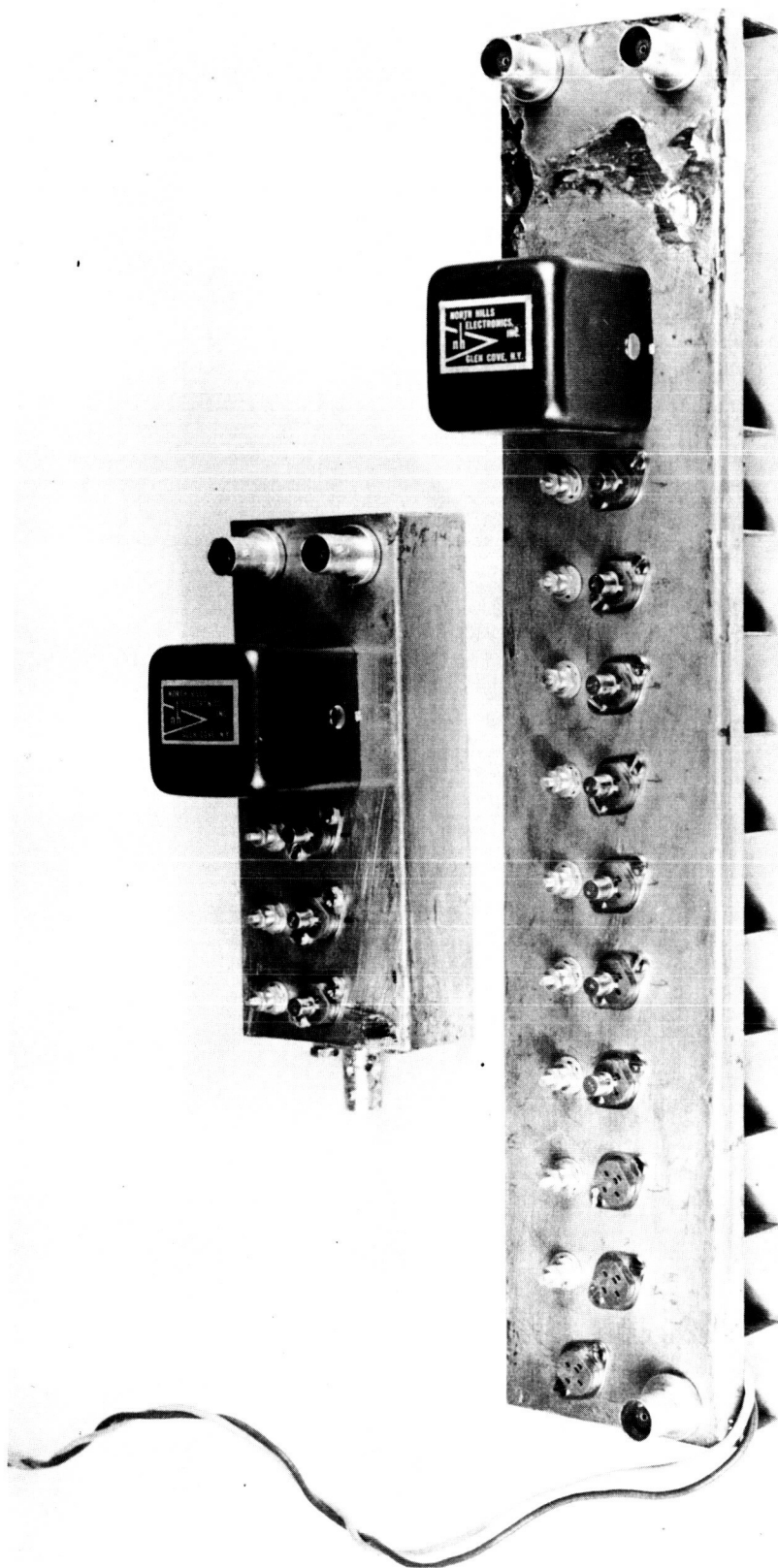
For a temperature of 290°K and a bandwidth of 100 mc, the noise voltage across a 50-ohm source resistor will be 8.9 microvolts rms. For a noiseless amplifier with a power gain of 80 db, the output voltage due to input noise voltage across the source resistance will be 89 millivolts. Additional noise from the transistors and any increase in bandwidth will then saturate the output stage of the amplifier. Consequently, the maximum power gain of an amplifier using the 2N2708 transistor with shunt feedback and a load

resistance of 50 ohms will be about 70 db.

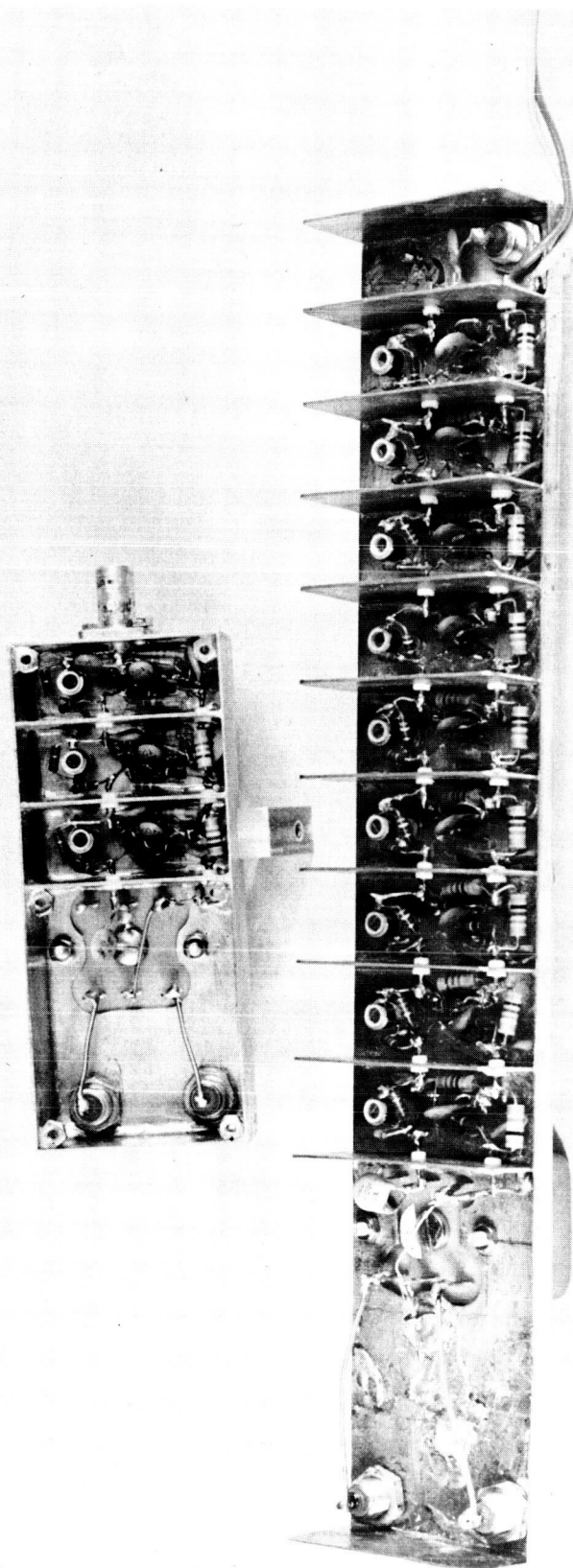
E. Seven-Stage Amplifier

Next, a seven-stage amplifier, with low-noise 2N2857 transistors in the first two stages and 2N2708 transistors in the remaining stages, was constructed. Photographs of the amplifier are shown in Figs. 11 and 12. A three-stage amplifier, also shown in the photographs, was originally constructed; but difficulty in working in the limited chassis space prompted the construction of a more spacious multi-stage amplifier chassis for experimental work. Consequently, data is not given for the three-stage amplifier. A circuit diagram of the seven stage amplifier is shown in Fig. 13. The transformer shown in the input circuit will be discussed in the next section.

Initially, the feedback network component values, given in Table 1-a, were adjusted to give a relatively flat frequency response over the 100-mc passband, as shown in Fig. 14. However, the noise figure of the amplifier was found to be excessive. Completely removing the feedback from the first stage of the amplifier considerably improved the noise figure, as indicated in Table 1-b. Unfortunately, the input impedance of the amplifier is no longer constant over the 100-mc passband. An attempt was made to measure the input admittance of the amplifier input stages, but the test equipment available was not capable of operating at power levels below the transistor input saturation point. However, a rough estimate of the input admittance of the amplifier can be made from y_{ie} of the 2N2857, given in Appendix 2.



TOP VIEW OF AMPLIFIERS
FIG. II.



BOTTOM VIEW OF AMPLIFIERS
FIG. 12.

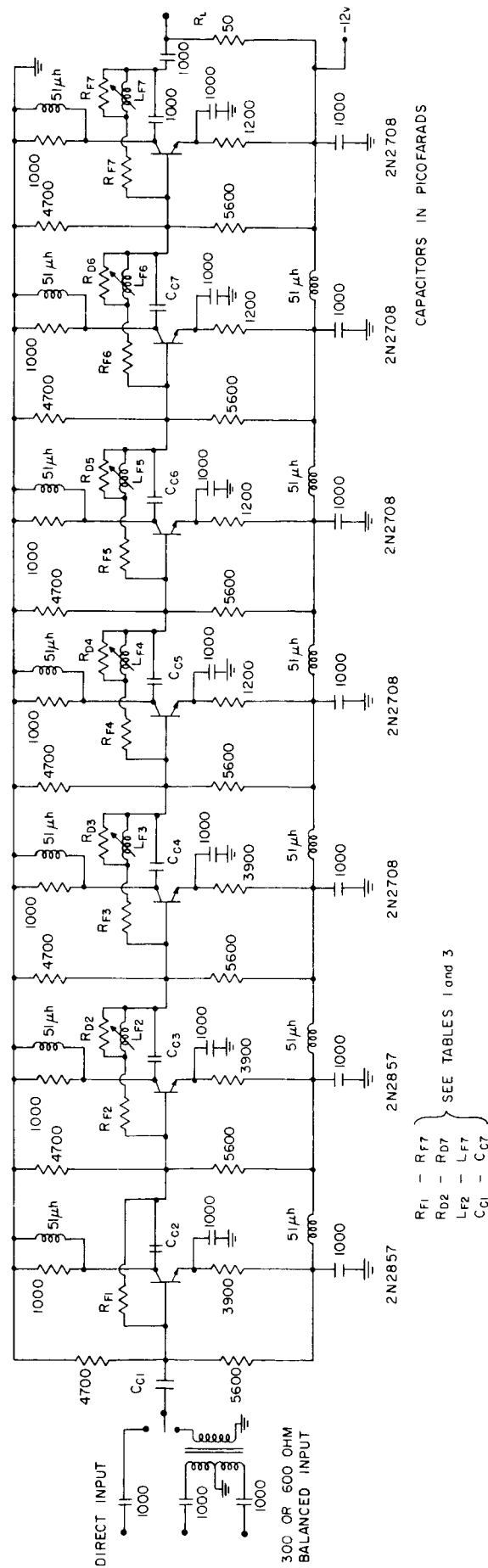


FIG. 13.

Stage	R_F (ohms)	R_D (ohms)	L_F (μ h)	C_c (pf)
1st	270 or ∞ (no feedback)	0	0	1000
2nd	270	180	.2-.3	1000
3rd	220	390	.3-.6	1000
4th	220	390	.2-.3	1000
5th	270	390	.4-.9	1000
6th	270	390	.3-.6	1000
7th	270	330	.4-.9	1000

Component Values

(a)

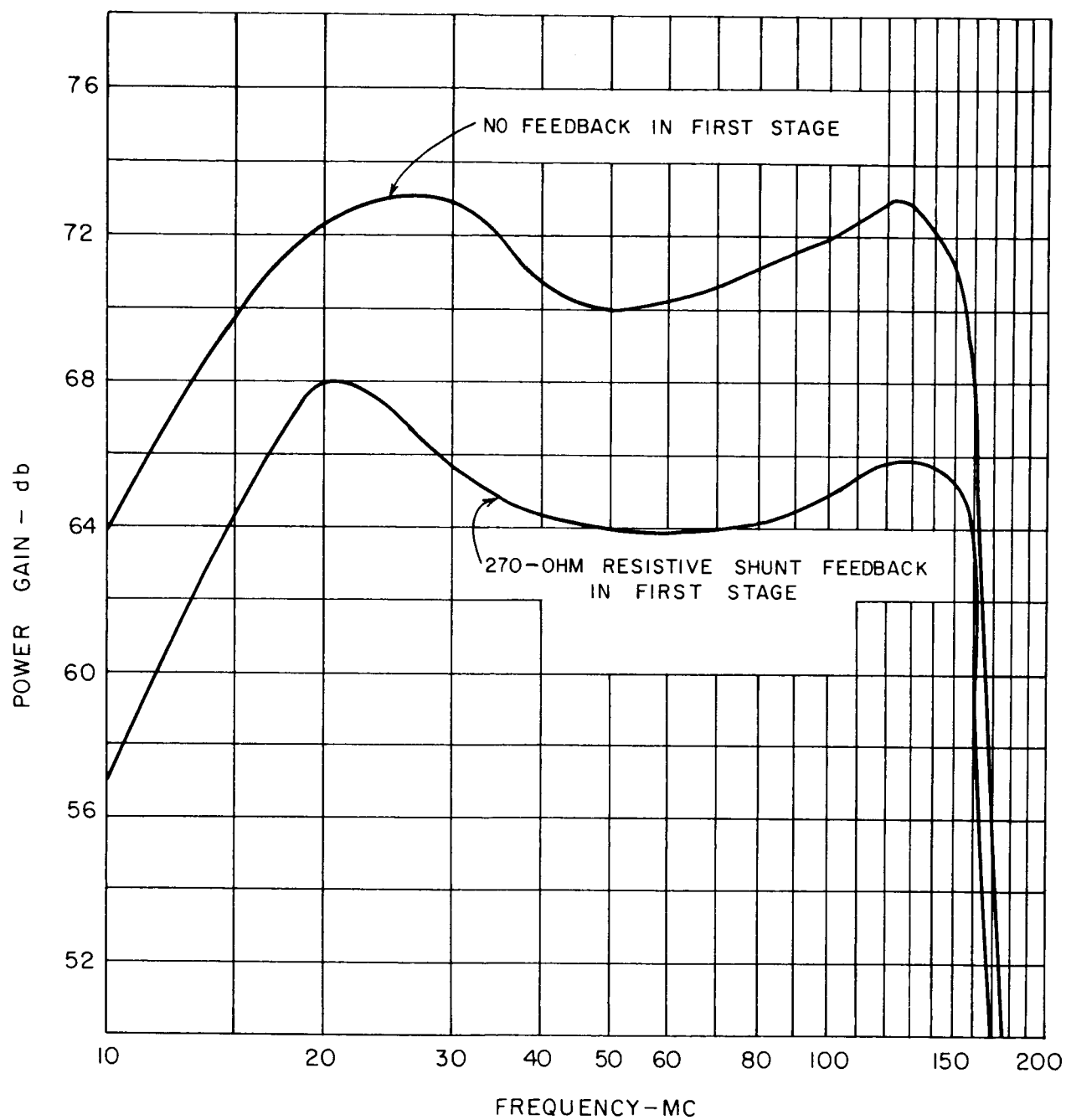
1st Stage, $R_F = 270$ ohms		1st Stage, $R_F = \infty$ (no feedback)	
Noise Generator Impedance(ohms)	Noise Figure (db)	Noise Generator Impedance(ohms)	Noise Figure (db)
50	5.2	50	3.0
75	5.8	75	3.3
150	6.4	150	3.5
300	8.0	300	3.8

Noise Figure, Direct Input

(b)

Seven-Stage Amplifier Component Values and Noise Figure

Table 1



FREQUENCY RESPONSE OF SEVEN-STAGE
AMPLIFIER; DIRECT INPUT

FIG. 14.

F. Impedance Matching for Balanced Input

Some wideband transformers were available for matching the 300-to 600 ohm impedance of the balanced mixer to the unbalanced input impedance of the transistor amplifier, approximately 70 ohms with a feedback resistance of 270 ohms. Two transformers, having impedance ratios of 600 ohms balanced to 75 ohms unbalanced and 300 ohms balanced to 75 ohms unbalanced, were obtained. These wideband transformers have very good impedance matching characteristics up to about 100 mc, as shown in the typical data sheets in Appendix 3. Since the shunt feedback amplifier stage has a wide range of gain control over the passband, the insertion loss falloff of the transformers above 100 mc can easily be equalized by adjusting the feedback network component values.

The frequency response of the seven-stage amplifier is shown in Fig. 15. One side of the balanced input was terminated with a 150-or 300-ohm resistor, depending on the transformer used, and the signal generator was applied to the other side. The noise figure of the amplifier with the balanced transformer input is given in Table 2. Again a considerable improvement in noise figure was apparent when the feedback was removed from the first stage.

G. 100-Mc Bandwidth Amplifier

An attempt was made to attenuate the seven-stage low-pass amplifier frequency response below 50 mc by inserting high-pass filters

1st Stage, $R_F = 270$ ohms		1st Stage, $R_F = \infty$ (no feedback)	
Noise Generator Impedance (ohms)	Noise Figure (db)	Noise Generator Impedance (ohms)	Noise Figure (db)
100	8.5	100	6.1
150	8.7	150	6.1
300	9.5	300	6.1
600	10.5	600	6.5

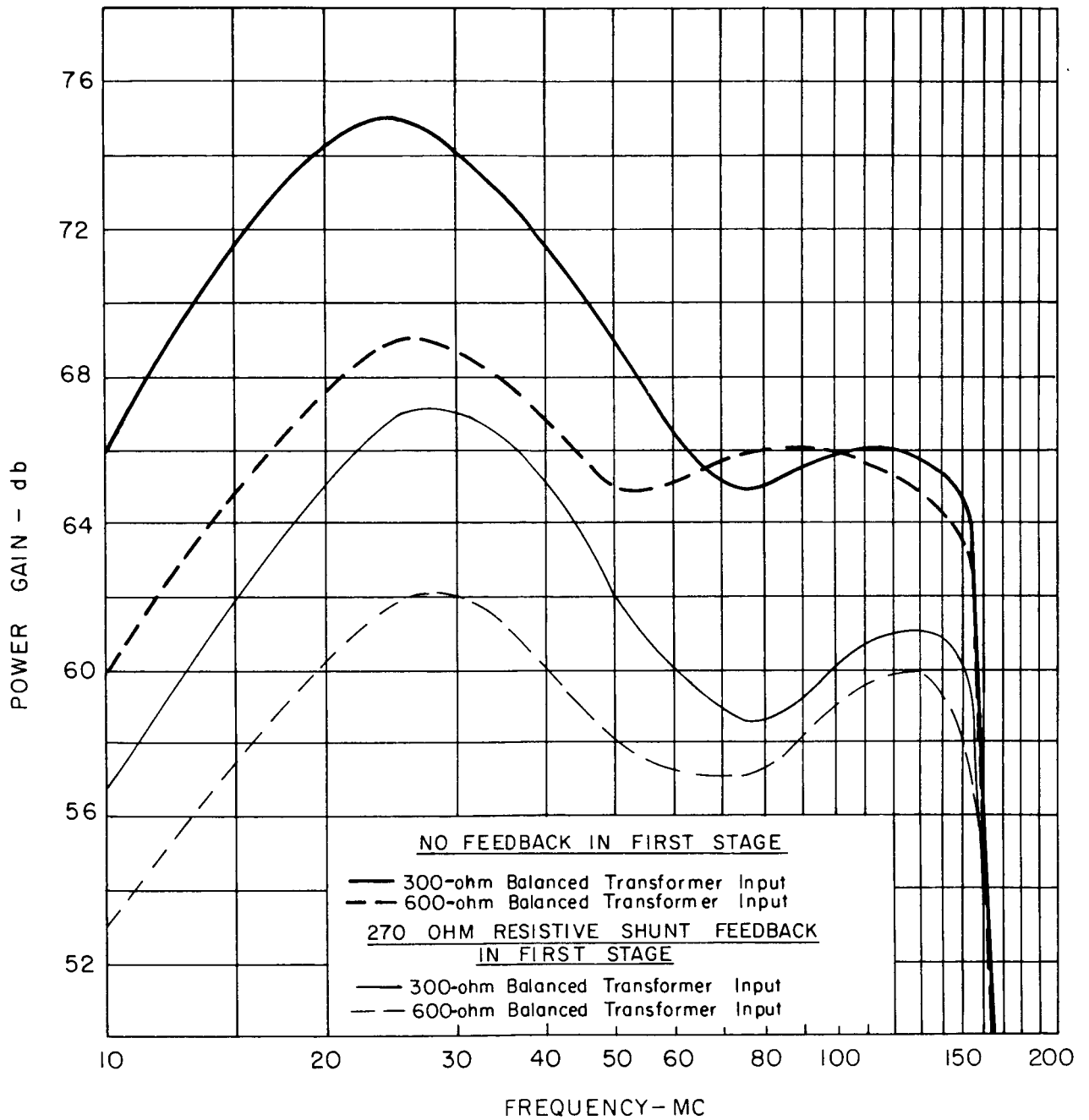
Noise Figure for 600-ohm Balanced Transformer Input
(a)

1st Stage, $R_F = 270$ ohms		1st Stage, $R_F = \infty$ (no feedback)	
Noise Generator Impedance (ohms)	Noise Figure (db)	Noise Generator Impedance (ohms)	Noise Figure (db)
100	6.7	100	4.8
150	7.4	150	4.8
300	8.0	300	4.8
600	9.5	600	5.3

Noise Figure for 300-ohm Balanced Transformer Input
(b)

Seven-Stage Amplifier Noise Figure; Balanced Transformer Input

Table 2



FREQUENCY RESPONSE OF SEVEN-STAGE AMPLIFIER;
BALANCED TRANSFORMER INPUT

FIG. 15.

in several stages, but proved unsuccessful because of filter and feedback reactance interaction. This reactance interaction seriously affected the amplifier frequency response over the 100-mc passband and could not be compensated for by adjusting the feedback network component values.

A much simpler method of attenuating the low-frequency response of the amplifier was tried. Decreasing the coupling capacitors, C_c , presented a high series impedance between stages at low frequencies, producing the required gain falloff below 50 mc. The low-frequency attenuation from the coupling capacitors was not as great as that of the high-pass filters, but adjustment of the feedback network component values to compensate for coupling capacitance interaction with the feedback reactance was possible. Values of the feedback network components are given in Table 3-a and frequency response of the amplifier is shown in Fig. 16. The increase in high-frequency gain necessary to compensate for insertion loss falloff of the wideband transformers above 100 mc was very evident in the frequency response of the amplifier with direct input, as shown in Fig. 16.

The noise figure of the 100-mc bandwidth amplifier, given in Table 3-b, is higher than the noise figure for the seven-stage low-pass amplifier. This is probably due to the improved noise figure of the transistors at lower frequencies and the pronounced gain peaking of the low-pass amplifier in the 25-mc region.

Stage	R_F (ohms)	R_D (ohms)	L_F (μ h)	C_c (pf)
1st	∞ (no feedback)	0	0	1000
2nd	270	470	.2-.3	15
3rd	270	680	.3-.6	47
4th	220	470	.3-.6	30
5th	270	390	.4-.9	1000
6th	270	390	.3-.6	25
7th	330	390	.3-.6	30

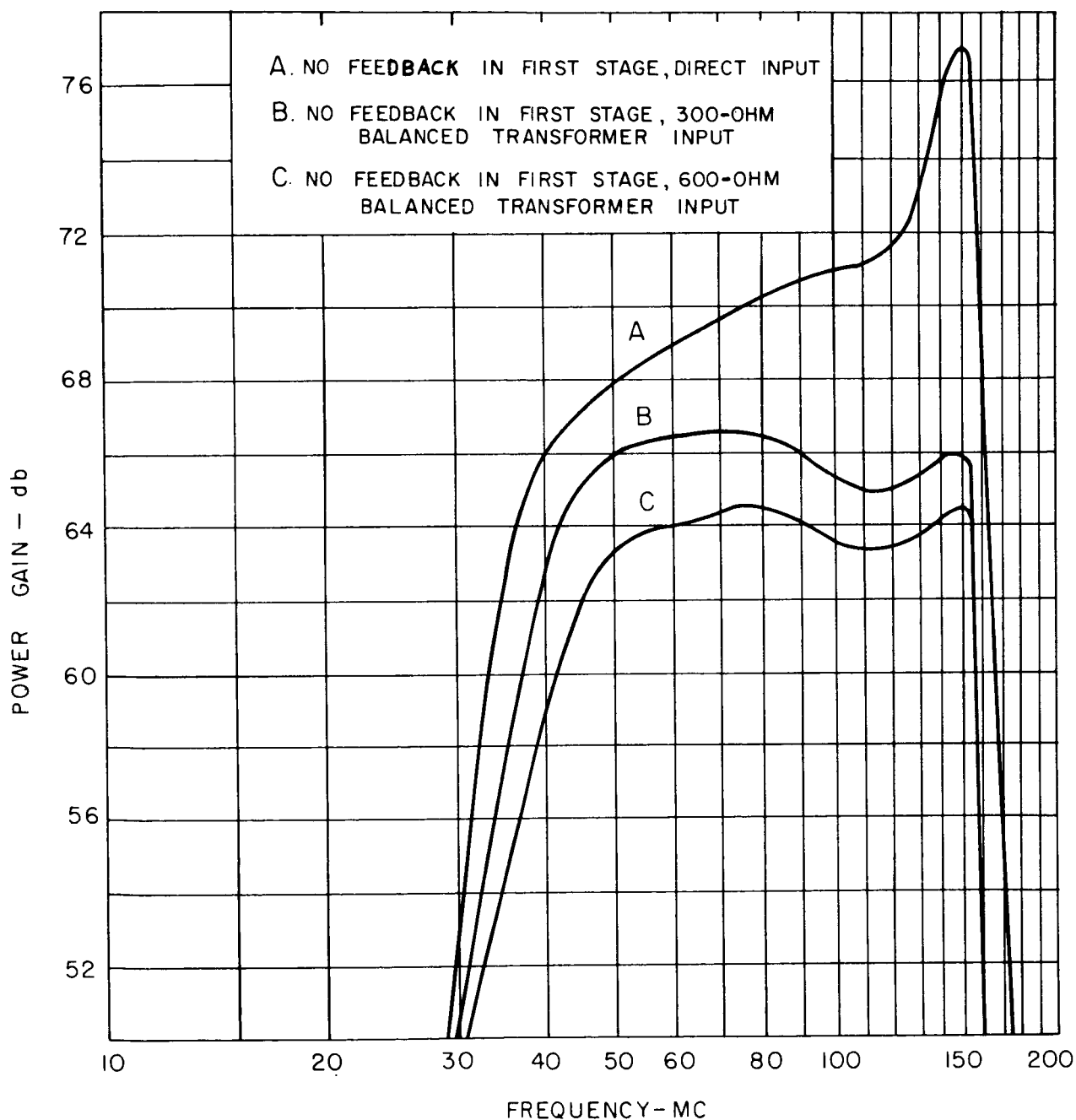
Component Values
(a)

600-ohm Balanced Transformer Input, No Feedback 1st Stage		300-ohm Balanced Transformer Input, No Feedback 1st Stage	
Noise Generator Impedance (ohms)	Noise Figure (db)	Noise Generator Impedance (ohms)	Noise Figure (db)
100	6.7	100	6.3
150	7.5	150	6.7
300	7.6	300	7.0
600	8.7	600	8.5

Noise Figure, Balanced Transformer Input
(b)

Component Values and Noise Figure for Seven-Stage
100-Mc Bandwidth Amplifier

Table 3



FREQUENCY RESPONSE OF SEVEN-STAGE
100-Mc BANDWIDTH AMPLIFIER

FIG. 16.

Time did not permit a thorough evaluation of the gain stability of the 100-mc bandwidth amplifier, but preliminary tests indicated that the gain variation due to a 4% supply voltage variation was about 1.5%. Future tests will be made on the temperature and time stability of the amplifier gain.

An additional 10-db gain, putting the total power gain of the amplifier at about 75 db, was desired. The output load impedance was increased to 1000 ohms in order to increase the output voltage swing; but, since the amplifier output impedance was low, considerable mismatching existed, decreasing the power gain of the amplifier at high frequencies. However, a common-base output stage, which would have a low input impedance and a high output impedance, could be added to the amplifier to provide both increased gain and output voltage. The addition of this output stage will be made in the future, along with an investigation of the mismatching between the wideband balanced transformers and the input impedance of the amplifier, with feedback removed from the first stage.

V. SUMMARY

The research described in this thesis concerns the design and construction of a low-noise 100-mc bandwidth transistorized amplifier suitable for research in radio astronomy. Several methods of obtaining large bandwidths in transistorized amplifiers were discussed. The common-emitter video amplifier configuration with frequency-dependent negative shunt feedback was chosen for the wideband amplifier.

The flexibility of the shunt feedback amplifier in exchanging gain and bandwidth was demonstrated in the section on experimental work. The maximum power gain of the shunt feedback amplifier was found to be about 70 db, due to the feedback-limited output voltage swing and saturation from noise voltage.

Wideband balanced to unbalanced transformers were used to match the amplifier to the 300-to 600-ohm balanced mixer impedance. Removing the feedback from the first stage of the amplifier resulted in a considerable improvement in noise figure, but at the expense of mismatching between the wideband transformer and the amplifier input impedance, which was no longer constant over the 100-mc passband.

Attenuation of frequencies below 50 mc in the wideband video amplifier was accomplished by adjustment of the coupling capacitors between stages. The seven-stage amplifier had a gain of about 65 db, flat to within 1.5 db over the 100-mc passband, and a noise figure of 7.0 db for a 300-ohm balanced input and 8.7 db for a 600-ohm balanced input.

Preliminary gain stability tests indicated that a 4% variation in supply voltage produced a 1.5% change in the power gain.

Future work on the wideband amplifier will include a thorough investigation of gain stability and mismatching effects between the balanced transformer and input impedance of the amplifier. Also, an attempt to increase the power gain by the addition of a common-base output stage will be made.

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APPENDIX 1

RCA RF TRANSISTORS

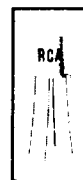


2N2708

RCA-2N2708 is a silicon n-p-n double-diffused epitaxial-type planar transistor intended for amplifier, mixer, and oscillator applications in the frequency range from 200 to 500 Mc.

The 2N2708 utilizes a 4-lead package that has the same case dimensions as the JEDEC TO-18 package.

SILICON N-P-N EPITAXIAL PLANAR TRANSISTOR



For VHF and UHF Applications

- high gain-bandwidth product (f_T)
700 Mc min.
- high max. neutralized power gain (G_{pe})
15 db min. at 200 Mc
- low collector-to-base time constant ($r_b C_c$)
33 psec max.
- high max. unneutralized power gain (G_{pe})
12 db typ. at 200 Mc

RF SERVICE

Maximum Ratings, Absolute-Maximum Values:

COLLECTOR-TO-BASE VOLTAGE, V_{CBO}	35	max.	volts
COLLECTOR-TO-EMITTER VOLTAGE, V_{CEO}	20	max.	volts
EMITTER-TO-BASE VOLTAGE, V_{EBO}	3	max.	volts
COLLECTOR CURRENT I_C	*	max.	ma
TRANSISTOR DISSIPATION:			
At free-air temperatures ---			
Up to 25° C	200	max.	mw
Above 25° C	Derate linearly 1.14 mw/°C		
TEMPERATURE:			
Storage	-65 to +200		°C
Operating (Junction).	200	max.	°C
Leads — At distances not less than 1/16"			
from seating surface for 10 sec. max.			
(during soldering).	230	max.	°C

* Limited by power dissipation.

Variation of Small-Signal "Y" Parameters (Typical) for Type 2N2708

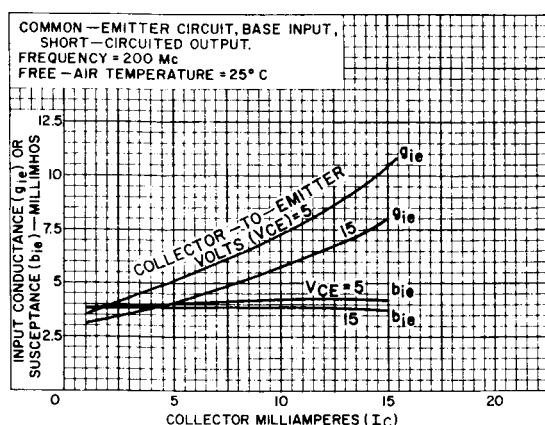


Fig. 1

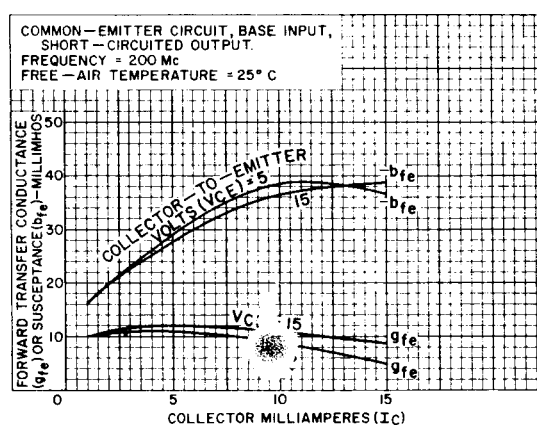


Fig. 2

ELECTRICAL CHARACTERISTICS, Free-Air Temperature = 25° C, Unless Otherwise Specified

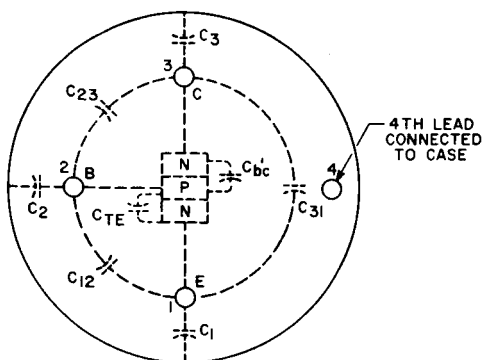
Characteristic	Symbol	TEST CONDITIONS						LIMITS		Units
		DC Collector Volts		DC Emitter Volts	DC Current (Milliamperes)			Type 2N2708		
		V _{CB}	V _{CE}	V _{EB}	I _E	I _B	I _C	Min.	Max.	
Collector-Cutoff Current: T _{FA} = 25° C 150° C	I _{CBO}	15 15			0 0			- -	0.01 1	μA μA
Collector-to-Base Breakdown Voltage	BV _{CBO}				0		1 μA	35	-	volts
Collector-to-Emitter Breakdown Voltage ^a	BV _{CEO} ^d					0	3	20	-	volts
DC Forward-Current Transfer Ratio	h _{FE}		2				2	30	200	
Magnitude of Small-Signal Forward-Current Transfer Ratio (Measured at 100 Mc)	h _{fe}		15				2	7	12	
Small-Signal Forward-Current Transfer Ratio (at f = 1 kc)	h _{fe}		15				2	30	180	
Emitter-to-Base Breakdown Voltage	BV _{EBO}				10 μA		0	3	-	volts
Output Capacitance (Measured at 140 Kc; see Fig. 3) ^b	C _{ob}	15			0			-	1.5	pf
Input Capacitance (Measured at 140 Kc; see Fig. 3) ^b	C _{ib}			0.5			0	1.4 (typ.)		pf
Collector-to-Base Time Constant (Measured at 31.9 Mc) ^c	r _b 'C _c	15					2	15	33	psec
Transconductance (Measured at 200 Mc)	g _m		15				2	25 (typ.)		mmho
Small-Signal Common Emitter Power Gain, Neutralized (Measured at 200 Mc; see Fig. 11) ^c	G _{pe}		15				2	15	22	db
Small-Signal Common Emitter Power Gain, Unneutralized (Measured at 200 Mc)	G _{pe}		15				2	12 (typ.)		db
Noise Figure: f = 200 Mc, source resistance = 50 ohms	NF		15				2	-	8.5	db

^a Pulse duration ≤ 300 μsec; duty factor ≤ 1%.

^b With lead No. 4 (case) not connected.

^c Lead No. 4 grounded.

^d Sustaining voltage.



92CS-11931

$$C_1 = C_2 = C_3 = 0.9 \text{ (approx.) pf}$$

$$C_{23} = 0.6 \text{ (approx.) pf}$$

$$C_{12} = 0.5 \text{ (approx.) pf}$$

$$C_{31} = 0.6 \text{ (approx.) pf}$$

C_{ob} = capacitance between base and collector (with collector reverse biased and emitter open)

With case ungrounded:

$$C_{ob} = C_2 C_3 / (C_2 + C_3) + C_{23} + C_{bc}'$$

Since C₂ = C₃,

$$C_{ob} = 0.9 \text{ pf} / 2 + 0.6 \text{ pf} + C_{bc}'$$

$$= 1.05 \text{ pf} + C_{bc}'$$

With case grounded:

$$C_{ob} = C_{23} + C_{bc}' + C_3$$

C_{ib} = capacitance between base and emitter (with emitter reverse biased and collector open)

With case ungrounded:

$$C_{ib} = C_2 C_1 / (C_2 + C_1) + C_2 + C_{TE}$$

$$= 0.95 \text{ pf} + C_{TE}$$

With case grounded:

$$C_{ib} = C_{12} + C_{TE} + C_2$$

C_{TE} = intrinsic base-to-emitter junction capacitance

C_{bc}' = intrinsic base-to-collector junction capacitance

Fig. 3 - Relationship of Capacitances C_{ob}, C_{ib}, C_{bc}', and C_{TE}.

Variation of Small-Signal "Y" Parameters (Typical) for Type 2N2708

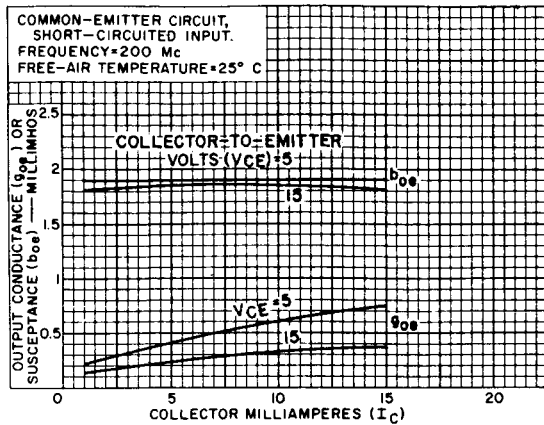


Fig. 4

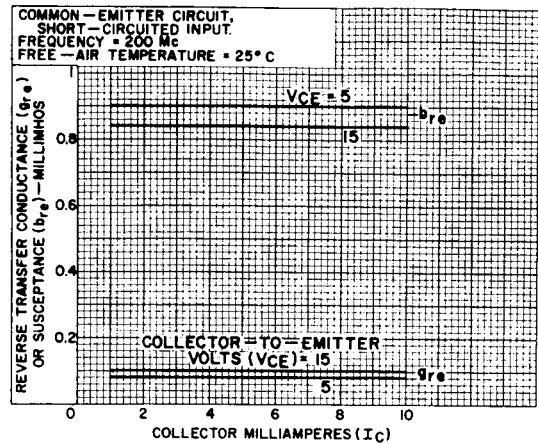


Fig. 5

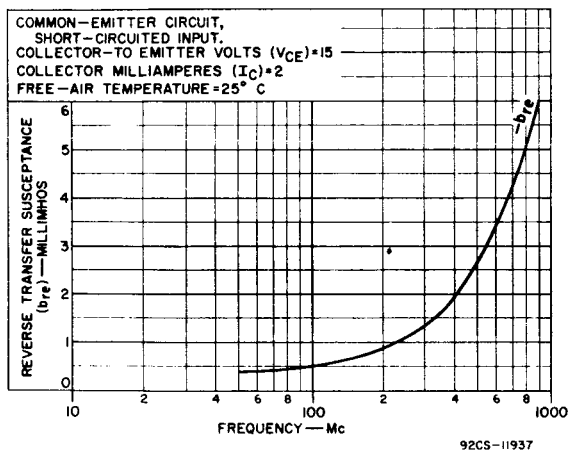


Fig. 6

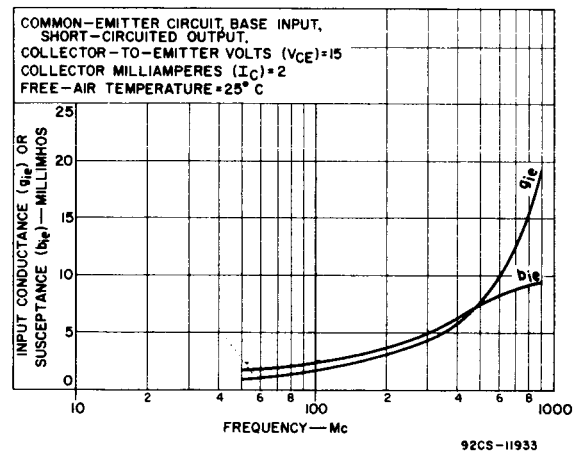


Fig. 7

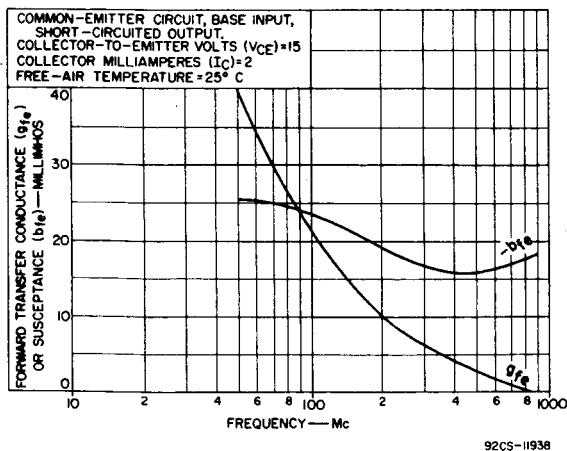


Fig. 8

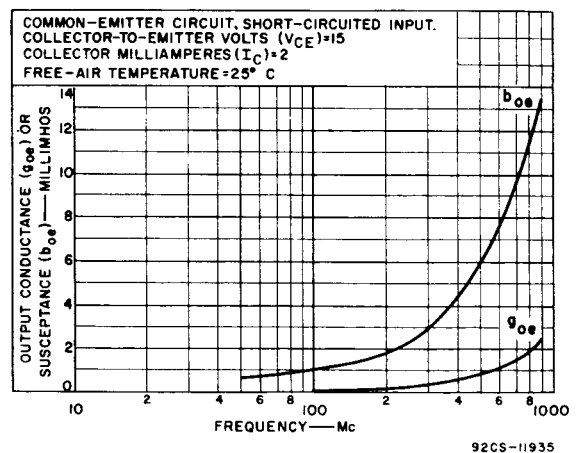
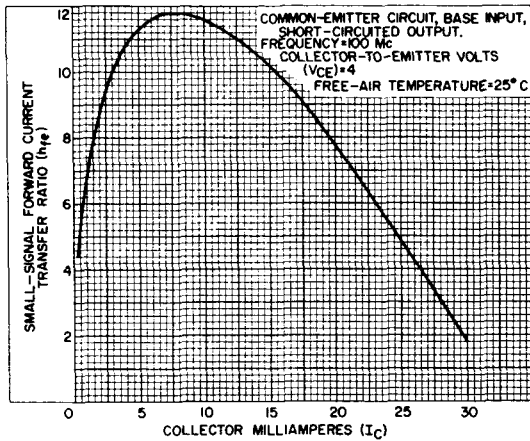


Fig. 9

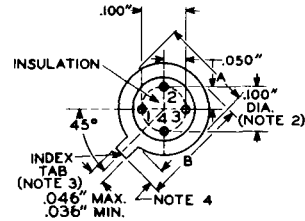
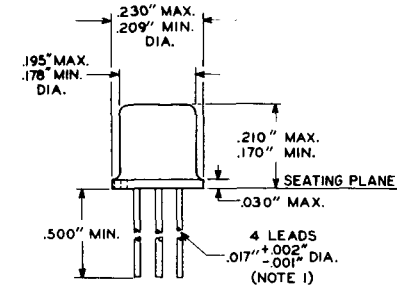
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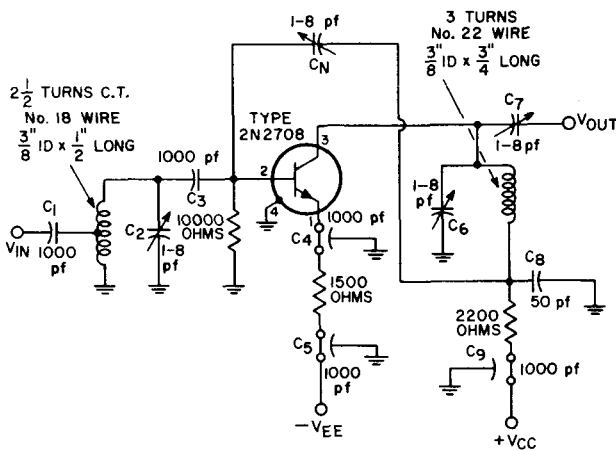
92CS-11940

Fig. 10 - Typical Small-Signal Forward-Current Transfer-Ratio Characteristic for 2N2708.

DIMENSIONAL OUTLINE



92CS-11941



92CS-11930

NOTE: (Neutralization Procedure): (a) Connect a 200-Mc signal generator (with $Z_{out} = 50$ ohms) to the input terminals of the amplifier. (b) Connect a 50-ohm r-f voltmeter across the output terminals of the amplifier. (c) Apply V_{EE} and V_{CC} , and with the signal generator adjusted for 10 mv output, tune C_2 , C_6 , and C_7 for maximum output. (d) Interchange the connections to the signal generator and the output indicator. (e) With sufficient signal applied to the output terminals of the amplifier, adjust C_N for a minimum indication at the input. (f) Repeat steps (a), (b), and (c) to determine if retuning is necessary.

Fig. 11 - Circuit of Neutralized Amplifier Used to Measure Power Gain at 200 Mc for Type 2N2708.

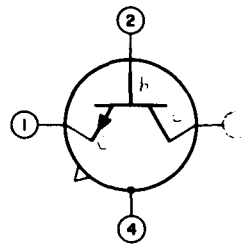
NOTE 1: THE SPECIFIED LEAD DIAMETER APPLIES IN THE ZONE BETWEEN 0.050" AND 0.250" FROM THE SEATING PLANE. FROM 0.250" TO THE END OF THE LEAD A MAXIMUM DIAMETER OF 0.021" IS HELD. OUTSIDE OF THESE ZONES, THE LEAD DIAMETER IS NOT CONTROLLED.

NOTE 2: MAXIMUM DIAMETER LEADS AT A GAUGING PLANE 0.054" \pm 0.001" - 0.000" BELOW SEATING PLANE TO BE WITHIN 0.007" OF THEIR TRUE LOCATION RELATIVE TO MAX. WIDTH TAB AND TO THE MAXIMUM 0.230" DIAMETER MEASURED WITH A SUITABLE GAUGE. WHEN GAUGE IS NOT USED, MEASUREMENT WILL BE MADE AT SEATING PLANE.

NOTE 3: FOR VISUAL ORIENTATION ONLY.

NOTE 4: TAB LENGTH TO BE 0.028" MINIMUM - 0.048" MAXIMUM, AND WILL BE DETERMINED BY SUBTRACTING DIAMETER A FROM DIMENSION B.

TERMINAL DIAGRAM Bottom View



LEAD 1 - EMITTER
LEAD 2 - BASE
LEAD 3 - COLLECTOR
LEAD 4 - CONNECTED TO CASE



RADIO CORPORATION OF AMERICA
SEMICONDUCTOR AND MATERIALS DIVISION
SOMERVILLE, N. J.

APPENDIX 2

RCA RF TRANSISTORS



2N2857

RCA-2N2857* is a double-diffused epitaxial planar transistor of the silicon n-p-n type. It is extremely useful in low-noise-amplifier, oscillator, and converter applications at frequencies up to 500 Mc with the common-emitter configuration, and up to 1200 Mc with the common-base configuration.

The 2N2857 utilizes a four-lead package which has the same case dimensions as the JEDEC TO-18 package. All active elements of the transistor are insulated from the case, which may be grounded by means of the fourth lead in applications requiring shielding of the device.

Maximum Ratings, Absolute-Maximum Values:

COLLECTOR-TO-BASE VOLTAGE, V_{CB0} . . .	30 max.	volts
COLLECTOR-TO-EMITTER VOLTAGE, V_{CE0} . .	15 max.	volts
EMITTER-TO-BASE VOLTAGE, V_{EB0}	2.5 max.	volts
COLLECTOR CURRENT, I_C	20 max.	ma

TRANSISTOR DISSIPATION, P_T :

For operation with heat sink:

At case temp. $\left\{ \begin{array}{l} \text{up to } 25^\circ \text{ C} \dots \dots \dots \end{array} \right.$	300 max.	mw
peratures $\left\{ \begin{array}{l} \text{above } 25^\circ \text{ C} \dots \dots \dots \end{array} \right.$	Derate at 1.72 mw/ $^\circ \text{C}$	

For operation in free air:

At free-air temp. $\left\{ \begin{array}{l} \text{up to } 25^\circ \text{ C} \dots \dots \dots \end{array} \right.$	200 max.	mw
temperatures $\left\{ \begin{array}{l} \text{above } 25^\circ \text{ C} \dots \dots \dots \end{array} \right.$	Derate at 1.14 mw/ $^\circ \text{C}$	

TEMPERATURE RANGE:

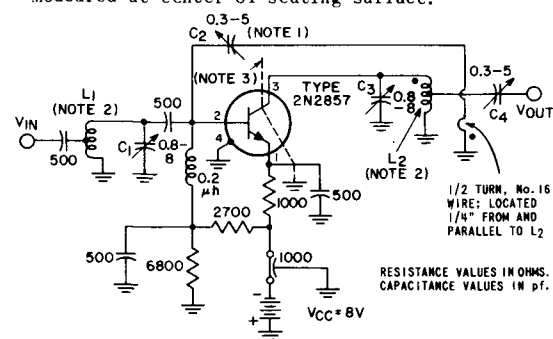
Storage and Operating (Junction) . . -65 to +200 $^\circ \text{C}$

LEAD TEMPERATURE (During soldering):

At distances $\leq 1/32$ inch from seating surface for 10 seconds max. 230 max. $^\circ \text{C}$

* Formerly Dev. No. TA-2333

** Measured at center of seating surface.



NOTE 1: (NEUTRALIZATION PROCEDURE): (A) CONNECT A 450-Mc SIGNAL GENERATOR (WITH $Z_{out} = 50$ OHMS) TO THE INPUT TERMINALS OF THE AMPLIFIER. (B) CONNECT A 50-OHM R-F VOLTMETER ACROSS THE OUTPUT TERMINALS OF THE AMPLIFIER. (C) APPLY V_{EE} AND V_{CC} , AND WITH THE SIGNAL GENERATOR ADJUSTED FOR 10 mv OUTPUT, TUNE C_1 , C_3 , AND C_4 FOR MAXIMUM OUTPUT. (D) INTERCHANGE THE CONNECTIONS TO THE SIGNAL GENERATOR AND THE OUTPUT INDICATOR. (E) WITH SUFFICIENT SIGNAL APPLIED TO THE OUTPUT TERMINALS OF THE AMPLIFIER, ADJUST C_2 FOR A MINIMUM INDICATION AT THE INPUT. (F) REPEAT STEPS (A), (B), AND (C) TO DETERMINE IF RETUNING IS NECESSARY.

NOTE 2: L_1 & L_2 — SILVER-PLATED BRASS ROD, 1-1/2" LONG \times 1/4" DIA. INSTALL AT LEAST 1/2" FROM NEAREST VERTICAL CHASSIS SURFACE. TAP 1-1/4" FROM CHASSIS END.

NOTE 3: EXTERNAL INTERLEAD SHIELD TO ISOLATE THE COLLECTOR LEAD FROM THE EMITTER AND BASE LEADS.

Fig. 1—Neutralized Amplifier Circuit Used to Measure 450-Mc Power Gain for Type 2N2857.

SILICON N-P-N EPITAXIAL PLANAR TRANSISTOR



For UHF Applications in Industrial and Military Equipment

Features:

- High Gain-Bandwidth Product—
 $f_T = 1000$ Mc min.
- High Converter (450-to-30 Mc) Gain—
 $G_c = 15$ db typ. for circuit bandwidth of approximately 2 Mc
- High Power Gain as Neutralized Amplifier—
 $G_{pe} = 12.5$ db min. at 450 Mc for circuit bandwidth of 20 Mc
- High Power Output as UHF Oscillator—
 $P_o = 30$ mw min., 40 mw typ. at 500 Mc
 $= 20$ mw typ. at 1 Gc
will oscillate at frequencies up to 2 Gc
- Low Device Noise Figure—
 $NF = 4.5$ db max. as 450-Mc Amplifier
 $= 7$ db typ. as 450-to-30 Mc Converter
- Low Collector-to-Base Time Constant—
 $r_b C_c = 15$ psec max.

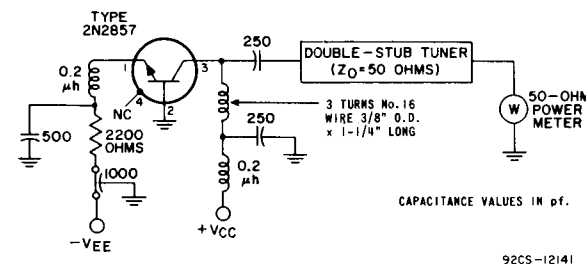
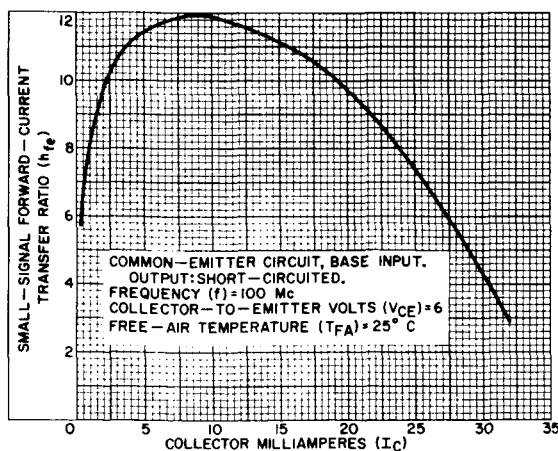


Fig. 2—Oscillator Circuit Used to Measure 500-Mc Power Output for Type 2N2857.

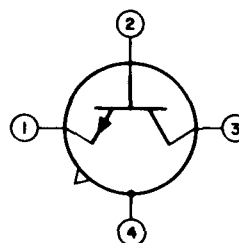
ELECTRICAL CHARACTERISTICS, At a Free-Air Temperature, T_{FA} , of 25° C:

Characteristic	Symbol	TEST CONDITIONS							LIMITS			Units
		Frequency f	DC Collector- to-Base Voltage V _{CB}	DC Collector- to-Emitter Voltage V _{CE}	DC Base-to- Emitter Voltage V _{BE}	DC Emitter Current I _E	DC Base Current I _B	DC Collec- tor Current I _C	Type 2N2857			
									Min.	Typ.	Max.	
Collector-Cutoff Current	I _{CBO}		15			0			-	-	0.01	μa
Collector-to-Base Breakdown Voltage	BV _{CB0}					0		0.001	30	-	-	volts
Collector-to-Emitter Breakdown Voltage	BV _{CE0}						0	3	15	-	-	volts
Emitter-to-Base Breakdown Voltage	BV _{EB0}					0.01		0	2.5	-	-	volts
DC Forward-Current Transfer Ratio	h _{FE}			1				3	30	-	150	
Small-Signal Forward- Current Transfer Ratio	h _{fe}	0.001 ^a 100 ^b		6 6				2 5	50 10	- -	220 19	
Output Capacitance	C _{ob}	0.140 ^a 0.140 ^b	10 10			0 0			- -	- -	1.8 1.3	pf pf
Input Capacitance	C _{ib}	0.140 ^a			0.5			0	-	1.4	-	pf
Collector-to-Base Time Constant	r _b ·C _c	31.9 ^b	6					2	4	-	15	psec
Small-Signal, Common- Emitter Power Gain in Neutralized Amplifier Circuit (See Fig.10)	G _{pe}	450 ^b		6				1.5	12.5	-	19	db
Power Output as Oscil- lator (See Fig.11)	P _o	500 ^b	10			-12			30	-	-	mw
UHF Noise Figure	NF	450 ^{b,c}		6				1.5	-	4	4.5	db

^a Fourth lead (case) not connected.^b Fourth lead (case) grounded.^c Source Resistance, R_g = 50 ohms.

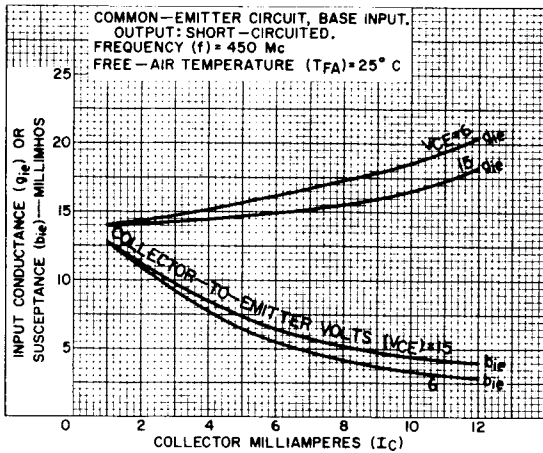
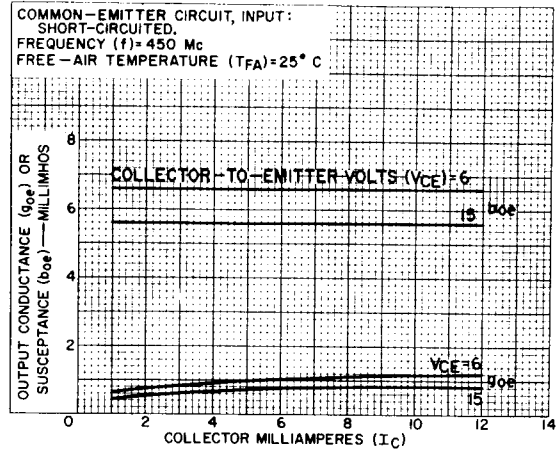
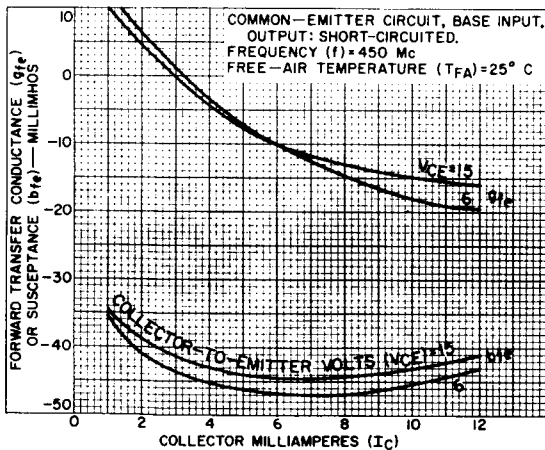
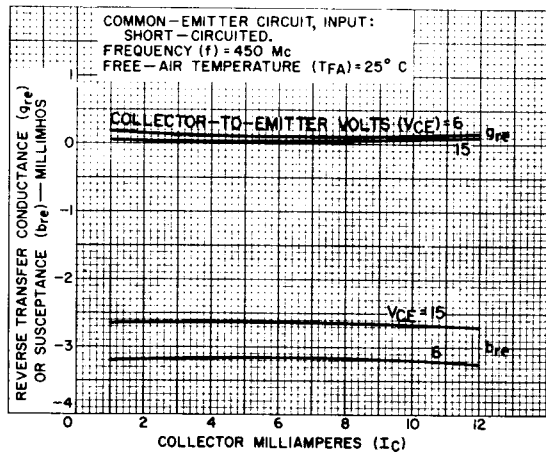
92CS-12153

Fig. 3 - Small-Signal Beta Characteristic for Type 2N2857.

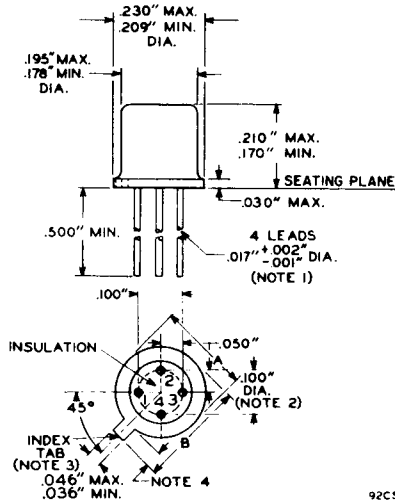
TERMINAL DIAGRAM
Bottom View

LEAD 1 - EMITTER
LEAD 2 - BASE
LEAD 3 - COLLECTOR
LEAD 4 - CONNECTED TO CASE

Two-Port Admittance (y) Parameters as Functions of Collector Current (I_C) for Type 2N2857

Fig. 4 - Input Admittance (y_{ie})Fig. 6 - Output Admittance (y_{oe})Fig. 5 - Forward Transadmittance (y_{fe})Fig. 7 - Reverse Transadmittance (y_{re})

DIMENSIONAL OUTLINE



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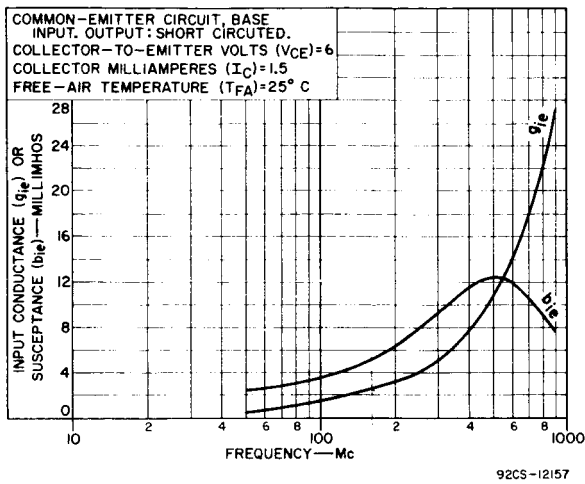
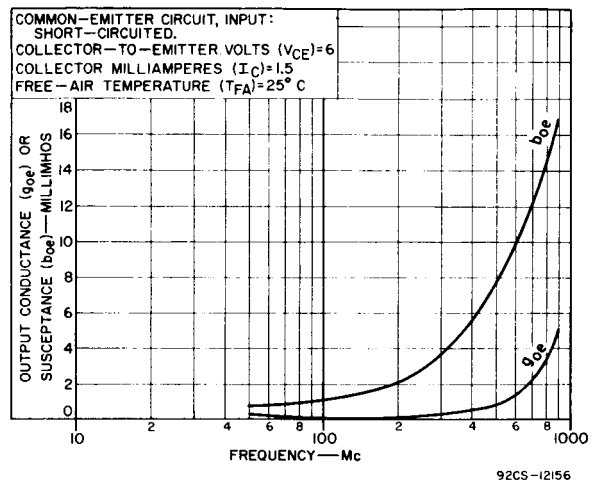
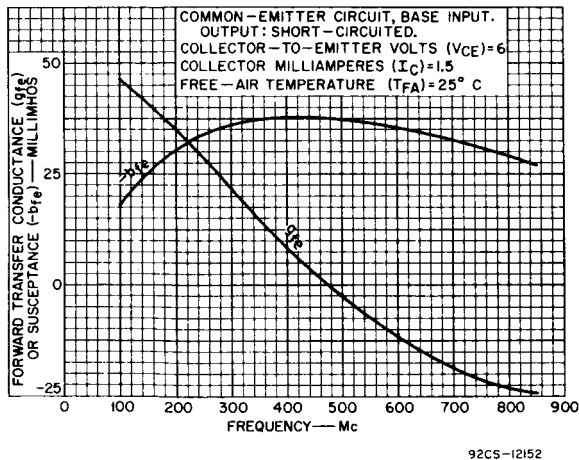
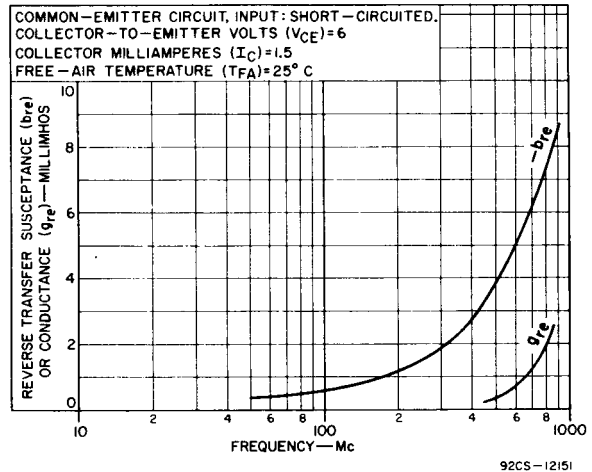
NOTE 1: THE SPECIFIED LEAD DIAMETER APPLIES IN THE ZONE BETWEEN 0.050" AND 0.250" FROM THE SEATING PLANE. FROM 0.250" TO THE END OF THE LEAD A MAXIMUM DIAMETER OF 0.021" IS HELD. OUTSIDE OF THESE ZONES, THE LEAD DIAMETER IS NOT CONTROLLED.

NOTE 2: MAXIMUM DIAMETER LEADS AT A GAUGING PLANE 0.054" + 0.001" - 0.000" BELOW SEATING PLANE TO BE WITHIN 0.007" OF THEIR TRUE LOCATION RELATIVE TO MAX. WIDTH TAB AND TO THE MAXIMUM 0.230" DIAMETER MEASURED WITH A SUITABLE GAUGE. WHEN GAUGE IS NOT USED, MEASUREMENT WILL BE MADE AT SEATING PLANE.

NOTE 3: FOR VISUAL ORIENTATION ONLY.

NOTE 4: TAB LENGTH TO BE 0.028" MINIMUM - 0.048" MAXIMUM, AND WILL BE DETERMINED BY SUBTRACTING DIAMETER A FROM DIMENSION B.

Two-Port Admittance (y) Parameters as Functions of Frequency (f) for Type 2N2857

Fig. 8 - Input Admittance (y_{ie})Fig. 10 - Output Admittance (y_{oe})Fig. 9 - Forward Transadmittance (y_{fe})Fig. 11 - Reverse Transadmittance (y_{re})

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RADIO CORPORATION OF AMERICA
ELECTRONIC COMPONENTS AND DEVICES

HARRISON, N.J.

APPENDIX 3

NORTH HILLS

WIDEBAND TRANSFORMERS

Application: Antenna Matching
Interstage Coupling
Impedance Matching
Computer Drive Circuits
Signal Distribution

Pulse Applications
Voltage Step-Up
DC Isolation
Phase Detectors
Balanced Modulators

Type No.	Impedance Ratio	Nom. BW MC	Type No.	Impedance Ratio	Nom. BW MC
0300	50 Unb/100 Bal	.1-100	1300	75 Unb/100 Bal	.1-125
0400	50 Unb/150 Bal	.1-100	1400	75 Unb/150 Bal	.1-65
0500	50 Unb/200 Bal	.1-100	1500	75 Unb/200 Bal	.1-125
0501	50 Unb/300 Bal	.1-100	1501	75 Unb/300 Bal	.1-100
* 0502			1600	75 Unb/450 Bal	.1-50
0600	50 Unb/450 Bal	.1-60	1700	75 Unb/600 Bal	.1-60
0700	50 Unb/600 Bal	.1-65	1701	75 Unb/600 Bal	.3-100
0800	50 Unb/800 Bal	.1-65	1702	75 Unb/600 Bal	.007-8
0900	50 Unb/1200 Bal	.1-30	1800	75 Unb/800 Bal	.2-30
1100	75 Unb/75 Bal	.1-125	1900	75 Unb/1200 Bal	.1-30
			1901	75 Unb/1850 Bal	.01-10
* 0502	50 Unb/200 Unb	.1-60	7700	600 Unb/600 Bal	.1-30

(Connector designed to mate with Hewlett-Packard 460A wideband amplifier). Series DB has BNC and DC has type N.

North Hills wideband transformers have been designed for low insertion loss (less than 1 db) and good matching characteristics over wide frequency ranges. Balanced windings feature extremely low unbalanced voltages and have a center tap. Data sheets with individual curves are available upon request.

Six case styles are offered. Series AA models are potted in anodized aluminum cases with a terminal board on the bottom and two 6-32 spade mounting bolts. Series BA are hermetically sealed with glass terminals at the case ends and two 4-40 studs on the bottom for mounting. Series BB is similar to BA except that the unbalanced connection is provided by a BNC connector. Series CA is a miniature plug-in unit. Provisions are made for plugging into a 7-pin socket or printed circuit board. This unit is 7/8" outside dimension by 1 1/2" long.

A great variety of different requirements is being met by special designs. Your specifications for custom wideband transformers may be submitted for quotation.

Prices

All Types AA & CA Case	0502 DB & DC Case	All Type BA & BB Case
Quantity	Quantity	Quantity
1-99 \$ 14.95	1-99 \$ 29.50	1-99 \$ 17.50
100-249 9.95	100-249 19.50	100-249 11.50
250-499 8.95	250-499 18.50	250-499 10.50
500-999 7.95	500-999 17.50	500-999 9.50

All prices subject to change without notice -
FOB Glen Cove, Long Island, New York
Terms 1/2% 10 days, net 30 days

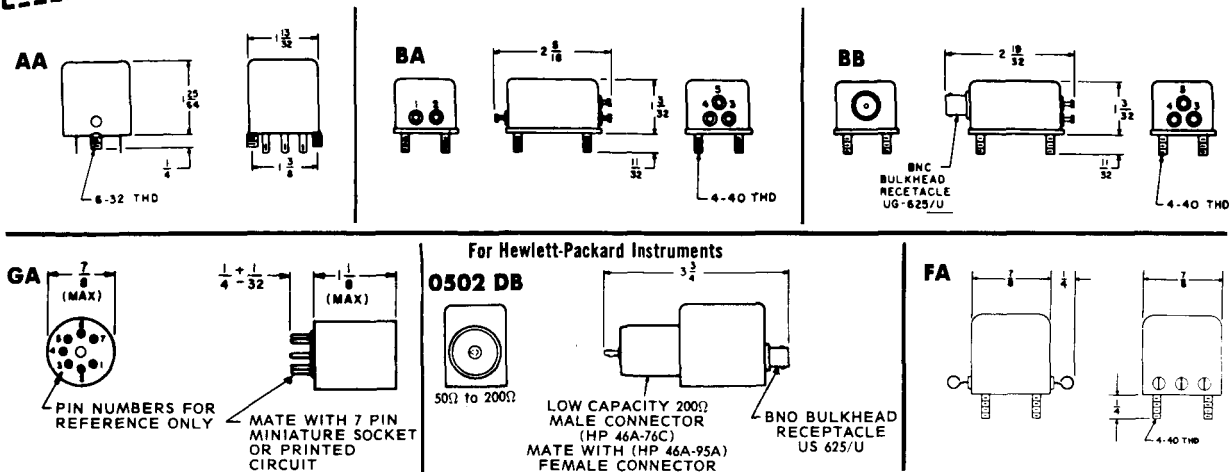
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ALEXANDER PLACE, GLEN COVE, NEW YORK



NORTH HILLS WIDE BAND TRANSFORMERS



- Antenna Matching
- Interstage Coupling
- Impedance Matching
- Computer Drive Circuits
- Signal Distribution
- Pulse Applications
- Voltage Step-Up
- DC Isolation
- Phase Detectors
- Balanced Modulators

Type Number	Impedance Ratio	Type Number	Impedance Ratio
1100	75 Unb/75 Bal	0100	50 Unb/100 Bal
1300	75 Unb/100 Bal	0300	50 Unb/100 Bal
1400	75 Unb/150 Bal	0400	50 Unb/150 Bal
1500	75 Unb/200 Bal	0500	50 Unb/200 Bal
1501	75 Unb/300 Bal	0501	50 Unb/300 Bal
1600	75 Unb/450 Bal	0600	50 Unb/450 Bal
1700	75 Unb/600 Bal	0700	50 Unb/600 Bal
1701	75 Unb/600 Bal	0800	50 Unb/800 Bal
1702	75 Unb/600 Bal	0900	50 Unb/1200 Bal
1703*	75 Unb/600 Bal		
1800	75 Unb/800 Bal		
1900	75 Unb/1200 Bal		
1901	75 Unb/1850 Bal		

* 10 watts

Data sheets for all types on request.

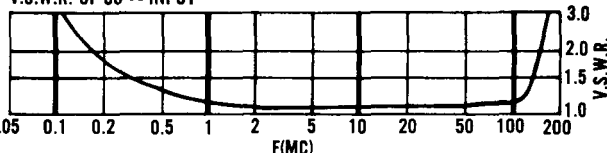
Frequency range on most of these units is contained within 100 kc to 100 mc band except for the 1702 and 1901 which cover from 20 kc to 8 mc. Data sheets with individual curves are available upon request.

North Hills wideband transformers have been designed for low insertion loss (less than 1 db) and good matching characteristics over wide frequency ranges. Balanced windings feature extremely low unbalanced voltages and have a center tap. Complete specifications are available upon request.

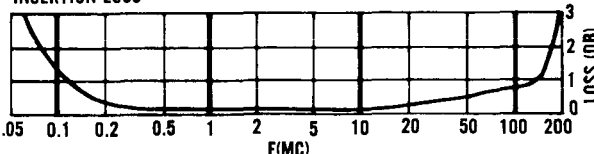
Several case styles are offered. Series AA models are potted in metal cases with a terminal board on the bottom and two 6-32 spade mounting bolts. Series BA are hermetically sealed with glass terminals at the case ends and two 4-40 studs on the bottom for mounting. Series BB is similar to BA except that the unbalanced connection is provided by a BNC connector. The above three styles are rated at 2 watts and the GA and the FA package are rated at 1/2 watt. Both GA and FA packages are hermetically sealed. Note that the types AA, BA and BB feature lower cut-off frequency than that of type GA and FA because of the larger core dimensions.

TYPICAL CHARACTERISTICS (0501)*

V.S.W.R. OF 50 Ω INPUT



INSERTION LOSS



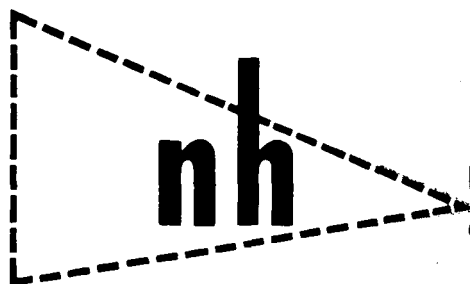
*Data sheets for all types on request.

Custom units designed to your specifications include power wideband transformers, pulse transformers, filters and adjustable inductors. See Radio Masters for standard line of inductors.

See Section 4000 for Precision AC and DC Power Sources.

See Section 2900 for Bridges, Potentiometers, Galvanometers, etc.

See Variable Inductors in Section 1800.



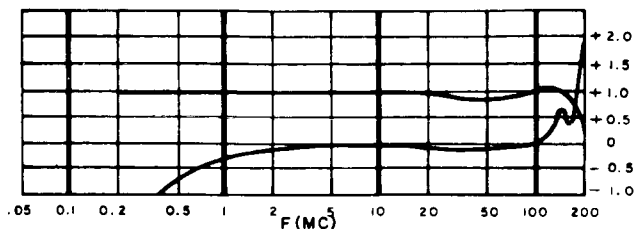
NORTH HILLS
ELECTRONICS, INCORPORATED
 GLEN COVE, LONG ISLAND, N. Y.

NORTH

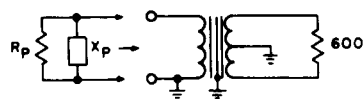
WIDEBAND TRANSFORMERS 1701 AA 75 Ω UNBAL : 600 Ω BAL BA BB

TYPICAL CHARACTERISTICS

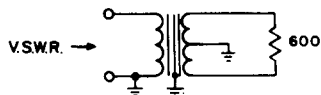
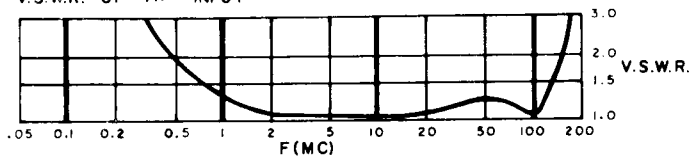
NORMALIZED IMPEDANCE OF 75 Ω INPUT



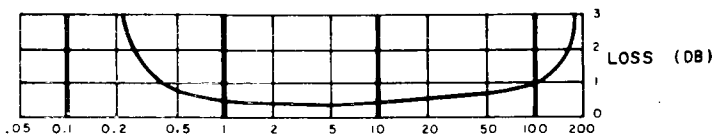
EQUIVALENT INPUT IMPEDANCE



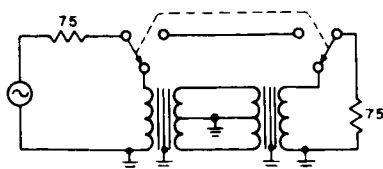
V.S.W.R. OF 75 Ω INPUT



INSERTION LOSS

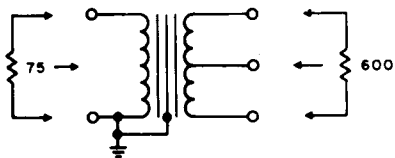


INSERTION LOSS $\approx \frac{1}{2}$ INSERTION LOSS OF TWO CASCADED TRANSFORMERS



PRI : 75 Ω UNBALANCED

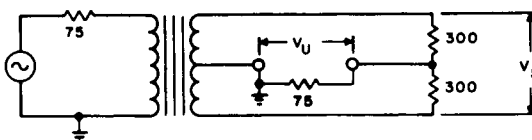
SEC : 600 Ω BALANCED WITH CENTER TAP



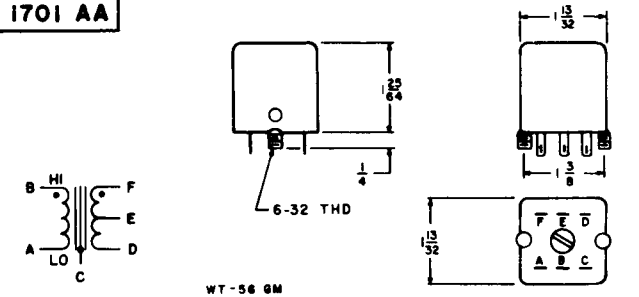
DIELECTRIC STRENGTH: 1000 V RMS TEST MAXIMUM
MAXIMUM POWER LEVEL : 1 WATT

SECONDARY BALANCE

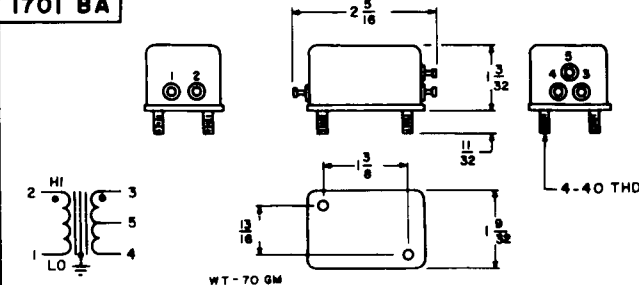
$\frac{V_S}{V_U} > 40$ DB AT 3 MC
 $\frac{V_S}{V_U} > 38$ DB AT 30 MC



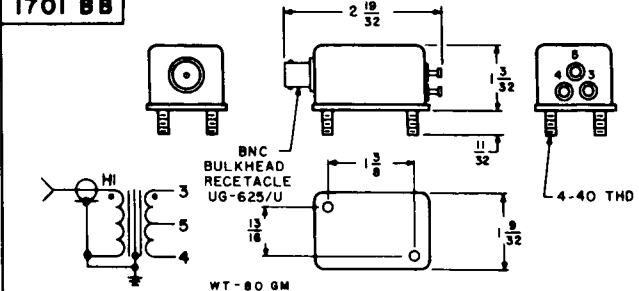
1701 AA



1701 BA



1701 BB



NORTH HILLS ELECTRONICS, INC.

GLEN COVE, L. I., N. Y. ORIOle 1-5700

BULLETIN NO.
1701

Lite in U.S.A.

VITA

Ronald A. Vivian was born in Detroit, Michigan, on March 7, 1940, and shortly thereafter moved to Carrizo Springs, Texas, with his parents, Mildred Vivian and the late Alfred W. Vivian.

After graduation from the Carrizo Springs Public Schools in 1958, he entered The University of Texas. In 1960, he transferred to the Massachusetts Institute of Technology, subsequently receiving the S.B. degree in Electrical Engineering in 1963.

In 1963, he returned to The University of Texas to continue his education, working for the degree of Master of Science in Electrical Engineering. Since February, 1964, he has held the position of Research Engineer with the Electrical Engineering Research Laboratory.

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Austin, Texas

This thesis was typed by Mrs. Maxine G. Krizov and Mrs. Mildred D. Thomas.